Proceedings of the I·R·E

DECEMBER 1942

VOLUME 30 NUMBER 12

Institute of Radio Engineers


42·S·17

Proceedings of the I·R·E

Published Monthly by
The Institute of Radio Engineers, Inc.

VOLUME 30 — *December, 1942* — NUMBER 12

Entered as second-class matter October 26, 1927, at the post office at Menasha, Wisconsin, under the Act of February 28, 1925, embodied in Paragraph 4, Section 538 of the Postal Laws and Regulations. Publication office, 450 Ahnaip Street, Menasha, Wisconsin. Editorial and advertising offices, 330 West 42nd St., New York, N. Y. Subscription $10.00 per year; foreign, $11.00.

The Institute of Radio Engineers
INCORPORATED

Winter Conferences, January 28, 1943, New York, N. Y.

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Preparation of Technical Articles*

BEVERLY DUDLEY†, ASSOCIATE, I.R.E.

Summary—*To appeal to the interests and intellectual requirements of the reader, the text of technical articles must be properly selected, arranged in logical order, and well proportioned, with essential concepts presented in proper relation to one another. To assist in meeting these requirements, a topical outline is given as a basis for the development of research or engineering articles. Recommendations are given for the typing, mailing, and proofreading of the manuscript. Suggestions are included for the preparation of diagrams and photographs so that they may be most suitable for publication.*

Introduction

TECHNICAL articles, regardless of their subject, are meant to represent a contribution to knowledge. For this reason alone, if not for many others, they should be prepared with utmost care. If they are written in a highly critical time, as that into which this decade falls, they should disseminate excellence in more ways than one. The English employed should receive careful consideration. Authors who pen technical articles are not expected to delight the connoisseurs of *belles lettres*. But it is not necessary that they completely neglect their English. A spoken word may be forgiven; it may also be forgotten. But a written word spells responsibility to its author at all times. If writers would pay more attention to their language, many a reader not only would receive a far clearer meaning of the subject conveyed, but would also derive greater enjoyment from the article.

Any article intended for publication in the technical press of the United States should be written in clear, concise, good English. There are many occasions where the author has failed to make his meaning clear, because he has underestimated this simple point. The author's meaning may be entirely clear to him, but if he would take the trouble to have someone else read his article many ambiguities might be corrected in the original manuscript.

Primary Purpose of Article

To be read is the primary purpose of any article. If it is eagerly read from beginning to end, the author reaps an additional reward for his name will have imprinted itself upon the mind of the reader, and articles bearing his by-line will be looked for. Any legitimate evocation which may induce the prospective reader to follow the text attentively is permissible, provided it is done within the definite limits of good taste. Good taste never should be extenuated and should serve as a guide for all technical publications because the writer will be regarded as an authority.

It is obvious that if the article is to be read it must be interesting to the group for which it is intended. A first consideration is that the subject matter must be appropriately selected. However, the manuscript must also appeal to the intellectual and academic standards of the readers.

General Considerations Determining the Type of Writing

The type of writing which should be employed will depend upon a number of factors. If the article is to be used for reference purposes, for example, each individual section of the article should be complete in itself so that the important information is readily apparent and immediately available. On the other hand, if the article is to be used for purposes of general instruction only, the requirement that each portion of the article be an independent unit need not be so rigorously followed.

The type of writing employed will depend to a very considerable extent upon the publication in which it is to appear. Therefore it is recommended that the author study the publication to which he intends to submit his manuscript in order that he may prepare his material in conformity with its style and policies. Very frequently such publications will have certain editorial and technical requirements which are published in the journal. By following these recommendations the author will save himself a considerable amount of time and effort in the proper preparation of his article.

Ethical and Censorship Considerations

The author has definite obligations to ascertain that the statements made are as precise and accurate as possible. Errors in the published article as well as inconsistencies and mistakes will reflect upon his technical ability.

At the present time each author submitting manuscripts for publication has an additional duty imposed upon him by virtue of the fact that the United States has entered into the world conflict. Much technical literature is potentially, if not actually, of such a nature as to be of aid or comfort to the enemy. Articles falling in this classification should not be published, or at least should be modified so that they conform to the best interests of the nation. It is recognized, of course, that by withholding the publication of technical or scientific information, the allies as well as the enemies may be handicapped and there appears to be some tendency to recognize this point of view. Censorship of technical articles is the responsibility of the individual author as well as that of the editor of the publication. If there is any doubt in the author's mind as to whether or not discussion of a particular phase of his work might be contrary to the interests of the nation, he

* Decimal classification: R050. Original manuscript received by the Institute, May 13, 1942. Presented, Winter Convention, New York, N. Y., January 13, 1942.

† Managing Editor, *Electronics*, McGraw-Hill Publishing Company, New York, N. Y.

should ascertain from his research director, or from the editor of the publication for which the article is intended, the advisability of publishing the results of his research.

Common sense demands that articles appearing in the technical press be prepared with good taste. Culture and ethics will dicate a limit to the number of times an author mentions a company's name in his article, and a reasonable degree of modesty in referring to the achievements of the persons in his organization. On the other hand, there have been cases where the work of others, who had directly or indirectly made contributions in the realm upon which the article touches, has been consciously or unconsciously overlooked. No author who submits an article for publication in one of the leading technical journals of the country should subject himself to such criticism for it is incumbent upon him to be familiar with the work of others. Fair play (if not cultural attainment) will demand that adequate recognition be given to the work of others, whether a part of his organization or not. Acknowledgment for assistance and similar courtesies may, with advantage, form a part of the published manuscript.

Five Important Points Which Help Make a Good Article

Only experience can be used to ascertain whether an article is a good one. No amount of written instructions can guarantee that an article composed in accordance with suggestions will be suitable throughout. Nevertheless, observance of the following important points will aid in assuring that the technical article is properly prepared.

(a) Selection. Include all material and data which may be necessary to bring out the main idea or message of the article. Reject all material which is irrelevant.

(b) Adaptation. The article should be adapted to the technical, cultural, academic, and intellectual attainment of the reader. As a corollary, the author should have in mind a particular type of reader for whom he is writing, even before he begins to prepare his manuscript.

(c) Arrangement. Arrange all parts of the article in a natural, logical order, keeping related topics together

(d) Articulation. The relationship of all parts of the article should be made clear and unmistakable. Much of this can be accomplished by clear, concise writing but considerable aid may also be obtained through the use of an appropriate outline in its preparation, or the use of heads and subheads in the manuscript itself to segregate appropriately the individual parts of the complete article.

(e) Proportion. The relative amount of space to be given to the various aspects of the subject should be determined by their relative importance to the article as a whole. The more important topics should receive proportionately more extensive treatment, while relatively little space should be devoted to insignificant details which the intelligent reader may be expected to supply for himself.

Writing the Manuscript

It is rather difficult to answer the question of optimum length of a manuscript since many factors enter into such a decision. The article should be long enough to treat all of the essential points adequately; no longer. The shorter an article is, while still conveying the same information, the more useful it will be. A good rule to follow in determining the length of an article is: Have something to say, say it, stop.

The author will have to determine the number of copies which should be prepared. In most cases the editor or publisher will require only one copy for his own use. But where the manuscript may be examined by several specialists in a particular field, the editor may require several copies. Publication in the Proceedings of the Institute of Radio Engineers will be expedited by submitting three copies of the manuscript and illustrations. In all cases the author should retain a copy of the manuscript for purposes of checking and proofreading.

An important point to consider in the initial preparation of an article (especially if it is mathematical) is the terms and symbolic notations employed. The author should be clear to specify at all times and in unmistakable form the terms and symbols to be employed, especially if they are used in some unusual or nonstandard form. It is advisable to employ terminology and symbolic notation which are universally recognized, and are in conformity with the standards of the various engineering and scientific societies.

No article should be written until all the necessary material which the author may require for its preparation is at hand. This will save lengthy interruptions and tend to minimize discontinuities of thought. It will be helpful if the article is planned in outline form before the text is begun. A written outline is usually most useful, but at least a mental picture of the article as a whole should preface the start.

The outline for a technical manuscript may include the following subjects:

Title
By-line
Summary or abstract
Introduction
Historical development of prior arts
Theory underlying the advances recorded in the article
Experimental method for verifying the theory
Description of experimental method and equipment
Results of experimental method
Analysis of experimental results
Correlation of theory and experiment
Account of the discrepancies between theory and experiment

Significance of discrepancies
Conclusion
Acknowledgment
Bibliography and references
Appendixes

The successful writer realizes the psychological values that an appropriate title holds. The right title is very important, yet few writers give this matter careful thought. In many cases the title sounds as if it had been tacked on hastily, just before sealing the envelope that is to carry the article to its final destination, as if the author had remembered at the last minute that the inclusion of a title was customary

The abstract or summary should be an abridgement of the essential material in the article. It enables the reader to obtain quickly an idea of the contents of the article so that he may determine whether these meet his present requirements in his search for information. The abstract must present as much information as possible as concisely as possible. The important results and conclusions should be indicated and the significance of the article always be clearly stated. It is not sufficient for the abstract or summary to state merely that certain research work has been undertaken since this conveys no information concerning the details of the research. It is much more desirable for the author to state his results and conclusions, and the significance of his work in relation to the present state of the science.

The introduction should be planned to acquaint the reader as thoroughly and as quickly as possible with the problems to come, their importance, their relation to other problems with which the reader is familiar and to pave the way for the main text which is to follow.

A sparkling introduction will immediately attract attention, but if the entire article is written in a brilliant style, it will hold the reader fascinated. Why should a technical article be unattractive? For too long a period we have seen in print uninspiring, clumsily written articles at the reading of which our eyes threatened to close several times, despite the undisputed technical contribution which they "elucidated." An article that fails to beguile us to reading it from start to finish has completely missed its aim. For a goodly number of years past we have reflected upon dullness as being commensurable with erudition; in reality we have been lacking imagination, originality, and pen appeal.

The conclusions should state clearly the deductions which are logically drawn from the treatment as developed in the article. The conclusion is mainly a statement of results achieved or new thoughts which have been derived as a result of the article.

Mechanical Considerations in Writing a Manuscript

The physical and mechanical form in which a manuscript is presented can do much to make an editor feel kindly or unkindly toward accepting an article.

It should go without saying that all copy should be typewritten with a machine having a clean ribbon and making a firm impression. All typing should be done on white paper $8\frac{1}{2}\times 11$ inches in size. It is especially desirable to avoid the use of legal-size paper since this does not conveniently fit into filing cabinets and the standard size of mailing envelopes. Margins of at least 1 inch should be left all around the paper although it is preferable to leave a margin of $1\frac{1}{2}$ inches on the left-hand side for binding.

All text should be double spaced, in order that appropriate editing and printer's marks may be inserted on the original manuscript.

Do not include drawings on pages of typed text. The text is sent to the printer who sets it in type whereas the drawings are usually sent to the engraver who makes cuts of them. If the drawings are on the same sheet with the text, it is necessary either to cut the drawings and text apart, or to have the printer and engraver work from the same sheet. Both practices are undesirable. The latter is especially bad since the copy may become easily damaged and soiled making it impossible for the engraver to produce satisfactory cuts from them.

Use of Mathematics

Sometimes the question arises as to the extent to which mathematics can be used advantageously in a technical article. The answer to this question can be given only by the author himself. He should remember that the use of an extensive amount of mathematical treatment will necessarily limit the audience which can appreciate his work since not all persons (not even all engineers and all scientists) are capable of reading and properly interpreting mathematical statements. Likewise, it should be recognized that mathematics are necessary only where quantitative considerations are involved. On the basis of these thoughts we may establish the following three points for the author to bear in mind: (1) Use mathematics only where they are needed to develop some quantitative relationships, (2) omit the use of mathematics where their employment is not essential for a clear understanding of the subject, (3) use only as much mathematics as are required, and keep these in the most simple form consistent with the development of the essential quantitative concepts.

An author can do much to simplify the mathematical form and complexity with which his article appears. This is especially true if he deals with relatively simple fractions. In such cases it is highly desirable, from a typographical point of view, to write such equations with the free use of parentheses, brackets, and fraction bars, rather than to use a long dividing line with numerator above and denominator below. This latter practice requires at least two or three lines of type and must be set by hand, whereas the recommended

practice frequently makes it possible to set a relatively complicated equation of the fraction type on a typesetting machine. Likewise, complicated exponents may be greatly shortened by writing it in the form "exp" followed by the exponent. For example, it is simpler to set the equation in the form

$$I = AT^2 \exp(-b_0/T)$$

than in the form

$$I = AT^2\epsilon^{-b_0/T}.$$

Typesetters are not mathematicians and the simpler the author can make the physical form of his mathematics, the more nearly certain he can be that his text will come through on galley or page proofs in the required manner.

Illustrating the Article

Illustrations are usually a highly important part of any well-prepared technical article. They may be used to present a better visual picture than can be done with words, to further the concept of quantitative relationship (as through the employment of graphs), or to create an easier understanding of the relationships to various parts than can be done in type.

All illustrations, whether photographs or line drawings, should be clear, clean cut, and to the point. The intended information should be designated easily and clearly and any extraneous material should be eliminated. In photographic illustrations, particular attention should be given to the background which may show material in which the author is not particularly interested. It is advisable to provide sufficiently wide margins in all cases. In making a cut for purposes of reproduction, the editor can always crop an illustration if it shows unnecessary extraneous matter, but he is not in a position to supply additional information which the author does not provide for him.

It is, of course, necessary to provide adequate captions for cuts for all illustrations. These captions together with their illustrations should be sufficiently complete to convey to the reader adequate information to enable him to interpret the illustration properly. It should not be necessary to read the text completely to understand what an illustration is about.

Requirements for Photographs

A good half-tone reproduction can be obtained only from a good photoengraving, and this in turn can only be made if the original photograph is a good one. Glossy 8-×10-inch photographic prints which are clear and sharp are by far the most suitable for publication. The "soot and whitewash" type of contrasty print is undesirable. The photographic print should show all tonal ranges from clear white to black, with good gradation in varying shades of gray. Cuts from which printed reproductions are made employ a metal plate whose surface has been finely screened and appreciable detail may be lost in the screening process. Consequently, the printed reproduction will always show less detail than the photographs from which it is made.

It is well to have photographs made by a professional photographer. Do not submit snapshots made with a small camera or in which the dominant subject is such a small part of the entire picture that it is lost. If you make the photographs yourself, use a camera making negatives at least 4×5 inches in size. If possible use a camera with a focusing back screen so that as the inverted image on the ground glass of the camera is viewed; the necessary camera adjustments and arrangement of equipment and lights can be made in the desired manner. In general, it is advisable to provide an adequate amount of diffuse illumination to make a relatively long exposure with a small aperture (corresponding to a large *f* number). By this procedure (which is of course suitable only for still objects) the sharpest and clearest of images may be obtained with the least amount of trouble and effort.

Legends should be provided for each photograph. Each must be complete and accurate to show its intended relationship to the article as a whole. The legends may be pasted on the back of the photographs or at the bottom of the prints. For the Proceedings it is preferred that they be typed together on a separate sheet and the illustrations marked on the border with figure numbers only. Whichever method be used, the legend must be unmistakably identified with the appropriate print. Do not attempt to write or type on the back of photographic prints as the impressions may show through on the glossy surface.

A photograph many times the size of the original should clearly indicate the amount or extent of magnification in the legend. Likewise, this magnification will have to be modified by the ratio of reduction when the engraver's cut is made from the print. It would be well for the author to check this change in size when he receives cut proofs from the editor.

Do not use paper clips on pictures as they frequently dent the picture's surface. Photographs should be mailed in plain envelopes with a cardboard stiffener to prevent or minimize damage.

Requirements for Sketches and Graphs

All drawings for reproduction must be in black on white paper, or preferably, on tracing cloth or Bristol board. Black India ink (with a carbon base) should be used in all cases.

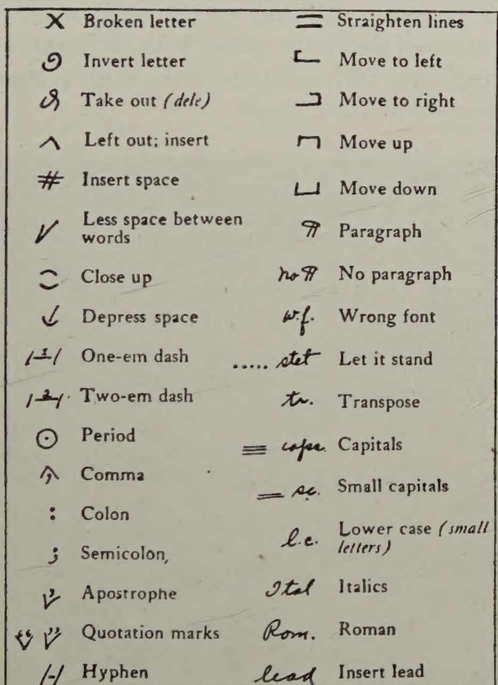
All graphs, drawings, and sketches should be in complete, finished form, properly lettered in ink so that the engraver's cut may be made directly from the drawing. Technical publications do not have the time, money, or facilities, nor can they assume the responsibility for having illustrations redrawn by a professional draftsman. It is the duty of the author to provide illustrations of sufficient quality that they may be used directly to make a half tone or line cut.

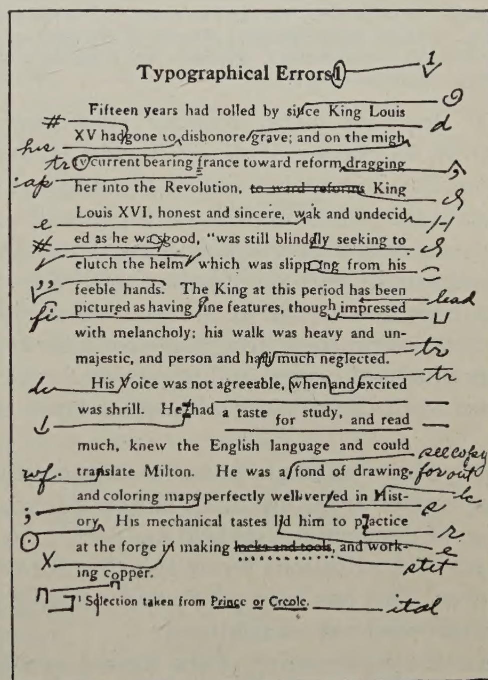
Lettering should be 1/4-inch high for graphs of

approximately 8×10 inches in size. Drawings which are smaller or larger than 8×10 inches should have lettering proportionately smaller or larger respectively.

Do not letter figure numbers or captions on the drawings. These should be inserted in the margin where they will not interfere with the sketch or graph.

Do not use orange or green cross-section paper for preparing graphs for reproduction. The co-ordinates invariably show up and produce an undesirable cross-hatch backing. If the divisions in the graph are close together, these lines run together and block up the drawing producing a dirty and untidy appearance. Since blue does not photograph on emulsions used by photoengravers, blue graph paper can be used and the blue co-ordinates will be eliminated when the engravings are made. All important co-ordinates should be properly drawn with India ink. The co-ordinates should be spaced at least 1/4 inch for an 8-×10-inch graphical illustration. Most engineering graphs or curves are simply used to indicate a variation of functional relationships. They are seldom intended for the reader to obtain exact values from the curves. However, if this is the purpose, then the graphs should be as large as possible and should be very accurately drawn.





Blueprints cannot be used for purposes of reproduction although they may be used for editorial examination. It is frequently convenient to make a tracing from which the original cuts may be made and to supply the editor with a number of additional blueprints (made from this tracing) which may be used in editing the article.

Transmitting the Article

It is advisable to send the manuscript by insured or registered mail. Insured mail enables the author to place an evaluation upon the manuscript and to collect a fee equal to this evaluation in the event the manuscript is lost. However, this is of little importance in most cases since one is hardly ever able to place a proper monetary evaluation on a manuscript. By sending the manuscript by registered mail, one is assured that everything humanly possible has been done to ascertain that the material reaches the person to whom it is directed.

Never fold any manuscript material or illustrations. If the illustrations are so large that they cannot easily be mailed flat, it is preferable to roll them up and send them through the mail in a cardboard mailing tube.

In submitting the manuscript it is advisable to accompany it by a letter of transmittal. This letter should indicate the title of the article, the author or authors, their affiliations, and to call attention to any matter which the author considers important or unusual. This letter of transmittal provides an excellent opportunity for the author to inquire about publishing procedure which may be unfamiliar to him and date of publication of his article. However, in the case of scientific or engineering organizations, where the manuscript must be read by others, it is not always possible to indicate definitely when an article may be published.

Proofreading

A manuscript which has been accepted for publication is marked for style and size of type, and is then set in type by the printer. The type is held in troughs or trays about 20 inches long and called galleys. "Galley proofs" are printed and returned for checking against the original manuscript. No new material should be added, nor should extensive changes be made

in these proofs. Such changes are always expensive and wasteful, but in technical publications, the addition of new material or changes should be particularly avoided, especially if there is possibility that the article may have important patent significance.

In correcting galley proofs, the more common standard proofreaders' marks, given here, will be found useful. Corrections should be indicated on the edge or border of the galley; they should not be written in the text itself.

He who passes judgment on himself will not be judged by others. If this admonition is observed in the preparation of technical articles, it may be safely assumed that the fate of the manuscript will be a happy one.

Copper-Oxide Rectifiers in Standard Broadcast Transmitters*

R. N. HARMON†, ASSOCIATE, I.R.E.

Summary—*Improvements in the processing of copper used in copper-oxide rectifiers make practical the use of this type of rectifier in modern broadcast transmitters. Features are reliability, long life, and ability to withstand surges.*

THE dry-type metal rectifier most commonly known as copper oxide or Rectox is, from an operating standpoint, one of the simplest pieces of apparatus for converting alternating into direct current. It is not only a very simple design, but also easy to operate; usually with outstanding reliability. Many attempts have been made in the past to develop this type of converter into an acceptable form for converting larger blocks of power, and to put the larger types on a more competitive basis with other kinds of converters.

Recently these attempts have been successful, and by giving proper attention to all factors entering into the design of this rectifier, its field has been enlarged to include many applications formerly using hot-cathode vapor tubes. This has been particularly true of recent commercial broadcast transmitters.

Because the construction of the Rectox metal rectifier is so very simple, not many ways have been found to increase its efficiency and output. Each rectifier element consists essentially of a copper disk with a surface layer of cuprous oxide (produced by oxidizing the surface of the disk) and of a second electrode that is in contact with the layer of cuprous oxide. The stack-type rectifier unit as now made consists of a number of such elements which, separated from one another by spacers and special cooling fins, are stacked on bolts and held together by end plates.

The most important material used in the manufacture of this rectifier is the copper. A thorough investigation of the properties of copper and the unavoidable impurities which it contains has revealed that, in general, copper from which certain impurities have been removed as completely as possible to be the most appropriate grade of copper for the manufacture of these rectifiers. Of great influence on the performance of the rectifiers is also the manner in which, beginning in the copper mill and ending in the annealing furnace, the metal is worked and treated.

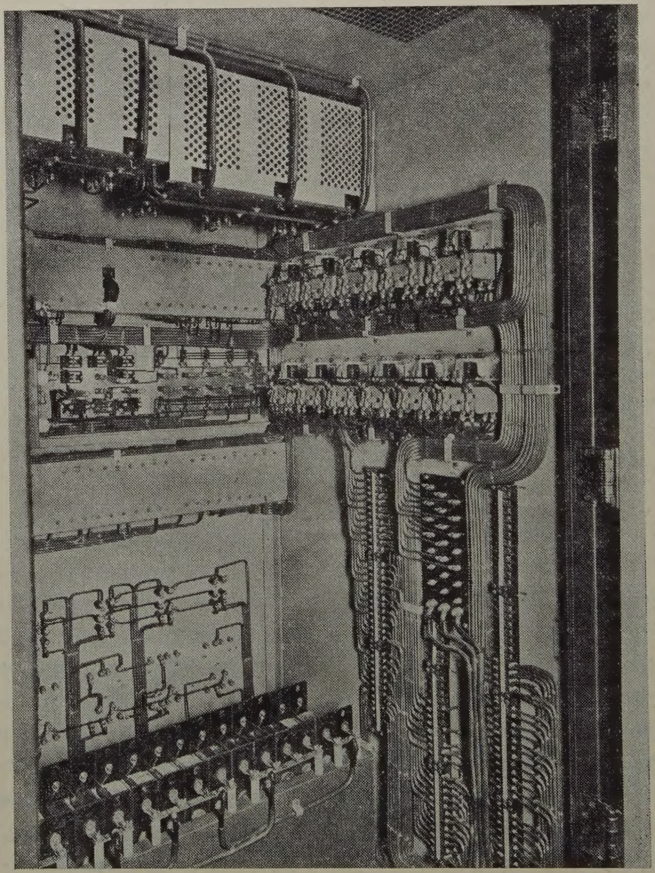


Fig. 1—Rear view of distribution cubicle for 50-kilowatt broadcast transmitter showing at bottom force-draft-cooled 3-phase full-wave 1250-volt, 0.7-ampere and 3-phase full-wave 3000-volt, 1.4-ampere rectifier for plate supplies.

By special annealing treatments, by careful choice of the copper, and by special production tests such rectifiers can be produced with great constancy, long life, and high efficiency. Further improvements in weight and size, as well as cost, have been made by forced-

* Decimal classification: R356.3. Original manuscript received by the Institute, May 13, 1942.

† Westinghouse Electric and Manufacturing Company, Baltimore, Maryland.

draft cooling which in some instances increased the output allowable more than ten times that obtainable with natural cooling.

The new forced-draft Rectox is well suited to application in broadcast transmitting apparatus. Outstanding advantages are its long life and reliability. No maintenance should be required. Even if a failure should occur, the plug-in construction permits replacement about as quickly as a tube. Its operation is instantaneous, that is, it is completely without time lag or any initial transient forming a build-up period. Hence the need for any time-delay device is eliminated, thus simplifying the control circuits.

In a standard line of 5-, 10-, and 50-kilowatt transmitters, all rectifiers except the main plate rectifiers are of this type having ratings from a few hundred volts and a quarter ampere up to 3000 volts and 1.4 amperes. Some of these units have been in operation over 15,000 hours. Nearly all of these units are forced-draft-cooled and are supplied air from the main cooling system used to cool the large radio tubes.

The Half-Wave Voltage-Doubling Rectifier Circuit*

D. L. WAIDELICH†, ASSOCIATE, I.R.E., AND C. H. GLEASON‡, ASSOCIATE, I.R.E.

Summary—*An analysis of the half-wave voltage-doubling rectifier circuit is made with the main assumption that the tube drop is zero while conducting. The performance characteristics of the circuit as predicted by the analysis are presented together with experimental verifications of several of these characteristics. Operating conditions for which polarized electrolytic condensers may be used and the currents to be expected on short circuit are discussed. The performance characteristics calculated from the analysis are presented as curves suitable for use in the prediction of the performance of an assembled circuit, and in the design of this doubler to meet specified operating conditions. A comparison is made of the performance characteristics of the half-wave and full-wave voltage doublers.*

Introduction

THE half-wave voltage doubler is being found useful as a power supply and as a component of high-voltage supplies. This circuit has several advantages over others employing input transformers. It offers economy in cost, size, and weight and hence is used in transformerless receivers. For use in radio-receiver power supplies, it has the important advantage of having a common input and output terminal.

Although no analysis of this half-wave doubler seems to have been made, several references to its operation and applications may be found.[1,2] Greinacher[3] seems to have made the first use of this circuit, employing it as the basic element of his voltage-multiplication circuit.

The results presented in this present paper on the half-wave doubler were obtained by a method of analysis similar to those employed in two previous analyses of the full-wave doubler.[4,5] The purposes of this paper are to present the results of the analysis by means of curves suitable for use in design, to compare some of the theoretical results with experimental results, and to compare the operating characteristics of the half-wave doubler with those of the full-wave doubler.

Analysis

The circuit diagram of the half-wave voltage doubler is shown in Fig. 1, and the current and voltage waveforms are shown in Fig. 2 for a complete cycle of the impressed alternating voltage e. Tube T_1 starts to conduct at $\omega t=\alpha$ and stops at $\omega t=\beta$, where $\omega/2\pi$ is the supply frequency and t is the time in seconds. Tube T_2 starts to conduct at $\omega t=\delta$ and stops at

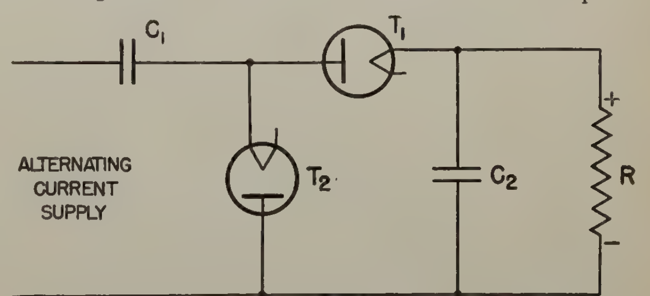


Fig. 1—Circuit diagram of the half-wave voltage-doubling rectifier.

$\omega t=-90$ degrees. Condenser C_1 is charged to approximately the peak value of the alternating voltage while tube T_2 is conducting and is discharged during the rest of the cycle. The voltage of C_1 is e_c and is shown in Fig. 2. The load voltage and also the voltage on C_2 have exactly the same shape as the load current i_L flowing through the load resistance R. The tube currents i_1 and i_2 have been reduced to one fifth of their size for convenience.

To simplify the analysis, the following assumptions are made: (1) the applied alternating voltage is sinusoidal, and the source has no impedance; (2) when conducting the tube drop is zero, and when not conducting the tube resistance is infinite; (3) the condensers have zero power factor and are both the same size;

* Decimal classification: R134×R366.3. Original manuscript received by the Institute, May 1, 1942. This paper constitutes a report on a project of the Engineering Experiment Station at the University of Missouri.

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‡ Westinghouse Electric and Manufacturing Company, Bloomfield, New Jersey.

[1] M. A. Honnell, "Applications of the voltage-doubler rectifier," *Communications*, vol. 20, p. 14; January, 1940.

[2] J. Millman and S. Seely, "Electronics," McGraw-Hill Book Company, New York, New York, 1941, p. 415.

[3] H. Greinacher, "Über eine Methode, Wechselstrom mittels elektrischer Ventile und Kondensatoren in hochgespannten Gleichstrom umzuwandeln," *Zeit. für Phys.*, vol. 4, pp. 195–205; February, 1921.

[4] D. L. Waidelich, "The full-wave voltage-doubling rectifier circuit," Proc. I.R.E., vol. 29, pp. 554–558; October, 1941.

[5] N. H. Roberts, "The diode as a half-wave, full-wave and voltage-doubling rectifier," *Wireless Eng.*, vol. 13, pp. 351–352; and pp. 423–430; July and August, 1936.

and (4) the load resistance has no inductance. The most serious of these assumptions is that of considering the tube drop to be zero, and this may be taken into account by an extension of this analysis.

Because the tubes act as open circuits when not conducting, the doubler circuit may be broken down into

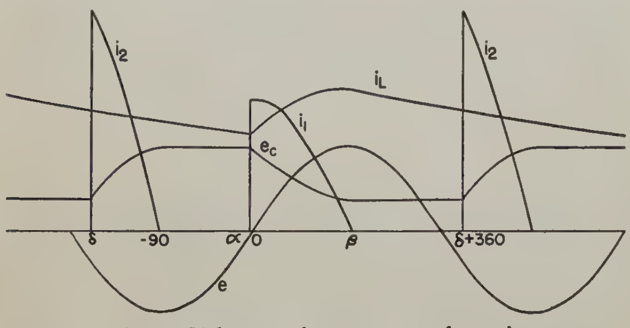

Fig. 2—Voltage and current waveforms in the half-wave voltage doubler.

three equivalent circuits which are in operation over certain portions of each cycle. While tube T_1 is conducting, from $\omega t=\alpha$ to $\omega t=\beta$, the equivalent circuit is Fig. 3(a). While neither tube is conducting, $\theta=\beta$ to $\omega t=(\delta+360$ degrees), the equivalent circuit is Fig. 3(b); and while tube T_2 is conducting, the doubler is represented by both Figs. 3(b) and 3(c). It can be shown that the tubes are never conducting at the same time, hence the equivalent circuits of Fig. 3(a) and 3(c) are never in operation simultaneously.

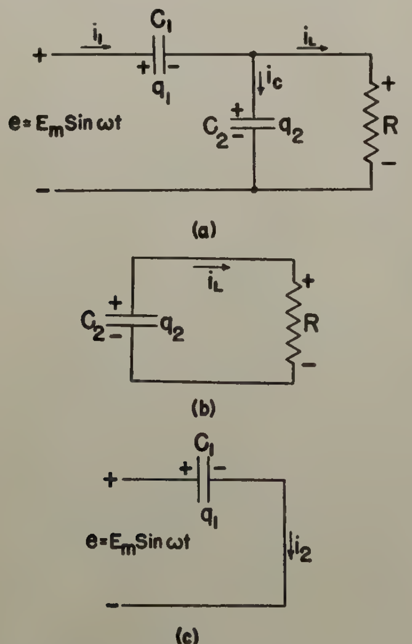

Fig. 3—(a) The equivalent circuit with tube T_1 conducting.
(b) The equivalent circuit with neither tube conducting.
(b) and (c) The equivalent circuits with tube T_2 conducting.

In spite of the simplifying assumptions made, the analysis is rather complex; therefore, only the analytical results and the more salient points of the analysis will be presented. An outline of the analysis may be found in Appendix II.

Before the voltage and current relations of the doubler can be found, the angles at which the tubes start and stop conducting must be determined. The analysis shows that all of these angles depend only upon the parameter (ωCR). In Fig. 4, for tube T_1, the angle α at which the tube starts conducting, the

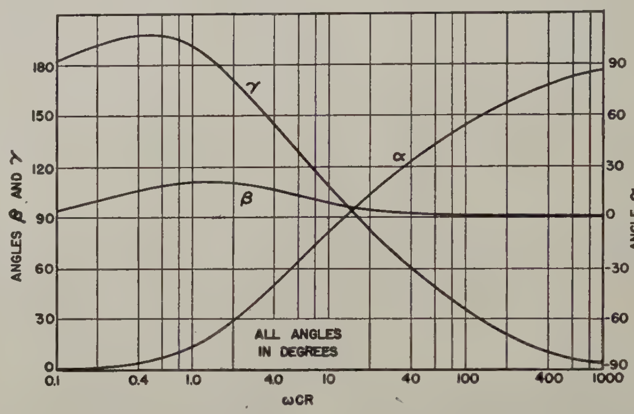

Fig. 4—The angles α, β, and γ of tube T_1.

angle β at which the tube stops conducting, and the angle γ during which the tube is conducting are shown as calculated for several values of (ωCR). For light loads (large values of ωCR) angles α and β approach 90 degrees and γ approaches zero degrees. As the load on the doubler is increased (ωCR decreasing), angle α approaches -90 degrees, and angle β rises to a maximum of 111 degrees and then decreases toward 90 degrees. The angle γ rises to a maximum of 195 degrees and then approaches 180 degrees as (ωCR) decreases toward zero.

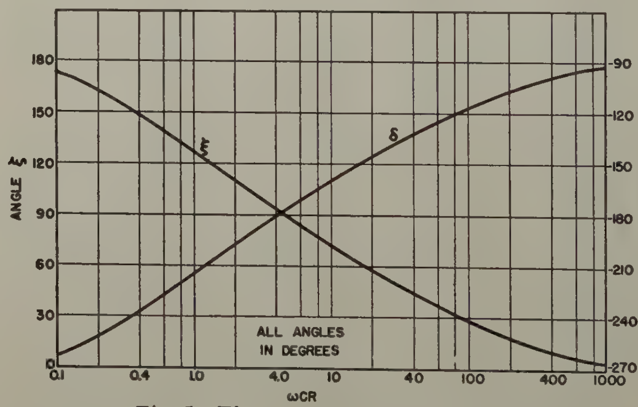

Fig. 5—The angles δ and ξ of tube T_2.

In Fig. 5 the angles of conduction for tube T_2 are also shown as functions of (ωCR) as calculated from the analysis. Angle δ at which the tube starts conducting is nearly -90 degrees for light loads and decreases toward -270 degrees as the load approaches short circuit. The angle ξ during which the tube is conducting varies from nearly zero degrees at light loads toward a limit of 180 degrees as the load is increased.

Characteristics

A most important characteristic of this circuit is the average output voltage for various loads. In Appendix II it is shown that the ratio of the average output

voltage to the maximum value of the impressed alternating voltage (E_{dc}/E_m) is a function of the parameter (ωCR) alone. In Fig. 6 this ratio is shown as a function of (ωCR) by a curve calculated from the analysis. For light loads this ratio is nearly two, dropping off toward zero for very heavy loads.

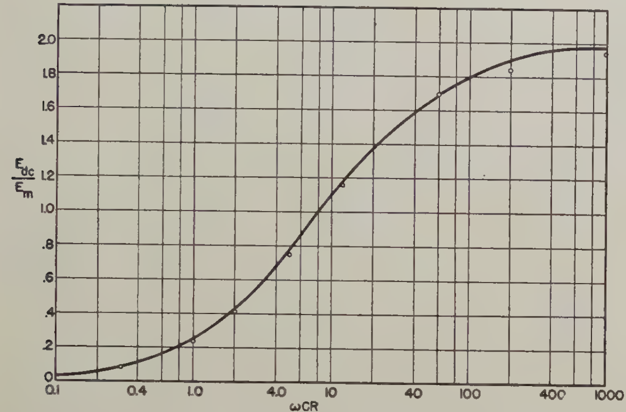

Fig. 6—The ratio of the output average voltage to the maximum value of the alternating supply voltage (E_{dc}/E_m). The circles are experimental points.

The output voltage contains some ripple voltage, with the fundamental frequency predominating. The per cent ripple r is defined as the per cent ratio of the effective ripple voltage to the average output voltage,[6] and is shown by the analysis to be solely a function of (ωCR). A curve of the per cent ripple r versus (ωCR) calculated from the analysis is shown in Fig. 7. At light loads (ωCR greater than 20) the per cent ripple is approximated by

$$r = 160/\omega CR.$$

The effect of the circuit upon the alternating-voltage supply is sometimes of interest and may be determined

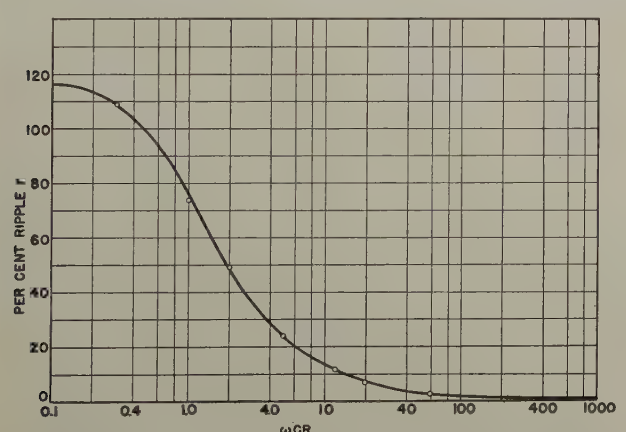

Fig. 7—The per cent ripple r. The experimental points are shown as circles.

from a knowledge of the effective current I and the input power factor. The ratio of the effective input current to the average output current (I/I_{dc}) and the power factor are also functions of (ωCR) only. Curves of these quantities calculated from the analysis are shown in Fig. 8. For light loads the tube currents are

[6] Institute of Radio Engineers, "Standards on Radio Receivers, 1938," p. 6.

quite peaked with a correspondingly high value of (I/I_{dc}). As the load approaches short circuit, this ratio approaches 2.22. The power factor is very low for high values of (ωCR) corresponding to the very peaked currents. It rises to a maximum of 54 per cent at $\omega CR = 30$ and decreases toward zero as the load approaches

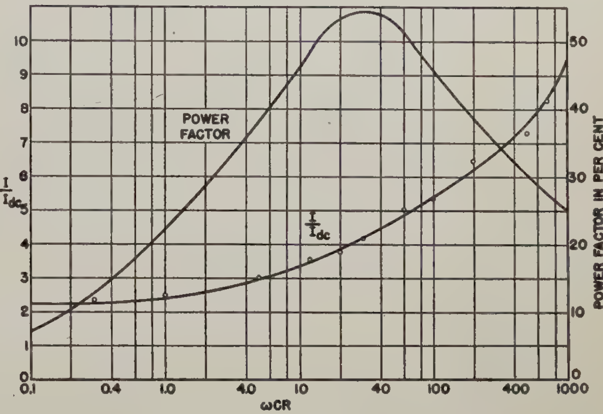

Fig. 8—The ratio of the effective input current to the average output current (I/I_{dc}) and the input power factor. The circles shown are experimental points.

short circuit, at which the doubler becomes a capacitive load on the alternating-voltage source. Examination of the doubler circuit shows that the average currents of each tube are the same and are equal to the average output current I_{dc}, which in turn may be found from Fig. 6. The ratio of the maximum tube current in tube T_1 to the average output current (i_{m1}/I_{dc}) is shown in Fig. 9 as a curve calculated from the analysis. Also shown is a similar curve for the ratio (i_{m2}/I_{dc}) for tube T_2.

The ratio of the peak inverse tube voltage to the maximum value of the applied voltage (e_p/E_m) is

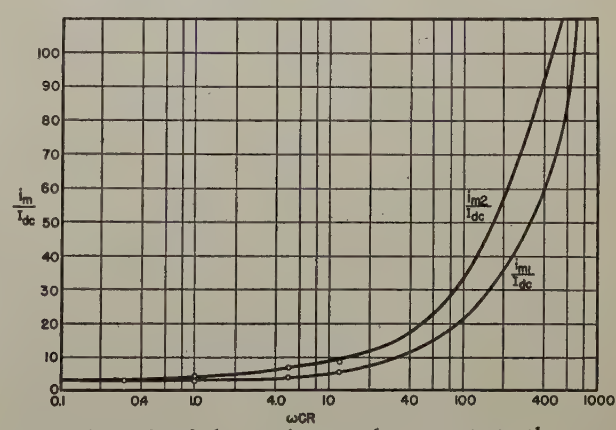

Fig. 9—The ratio of the maximum tube currents to the average output current (i_m/I_{dc}) for both tubes. The circles are experimental points.

shown in Fig. 10 for both tubes. These ratios are very nearly two at light loads and decrease toward zero as the load approaches short circuit.

As a verification of these calculated curves of the doubler's characteristics, some experimental results are shown for comparison. The ripple voltage was evaluated experimentally by measurement of the harmonic components with the aid of a wave analyzer. The

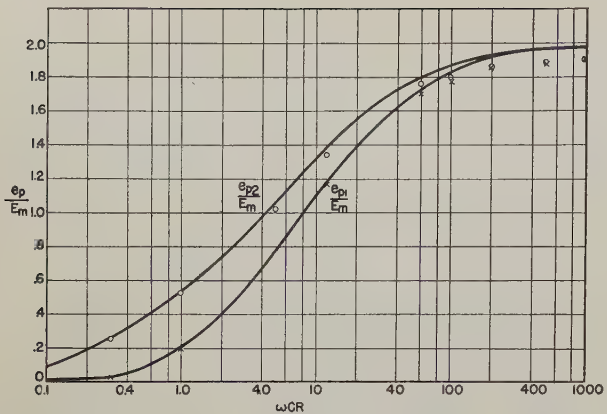

Fig. 10—The ratio of the maximum inverse tube voltage to the maximum value of the alternating supply voltage (e_p/E_m) for both tubes with experimental points shown.

maximum tube currents were measured by the use of a cathode-ray oscilloscope. The peak inverse tube voltages were measured with a peak-voltage voltmeter designed for this purpose.

Condensers

The input condenser C_1 must withstand the maximum value of the alternating-voltage supply E_m, and the load condenser C_2 must withstand a maximum voltage of $2E_m$. The analysis shows that the voltage on condenser C_1 reverses over a part of the cycle if the doubler is loaded heavily enough. This reversal occurs if the angle δ is less than -180 degrees, and from Fig. 5 this corresponds to values of (ωCR) less than 4.50. Hence a polarized electrolytic condenser may be used for C_1 only if (ωCR) is greater than 4.50. From Fig. 6 $(E_{dc}/E_m)=0.772$ for $(\omega CR)=4.50$; therefore, this value of (E_{dc}/E_m) may be used to determine whether or not this type of condenser may be used. This value of (E_{dc}/E_m) is below the usual operating values for this circuit, and hence in most practical circuits, polarized electrolytic condensers may be used without danger. In any case a polarized electrolytic condenser may be used for condenser C_2.

Operation on Short Circuit

When the doubler is short-circuited, the tubes conduct throughout alternate half cycles. The analysis shows that the effective input current on short circuit is

$$I = \omega C E_m/\sqrt{2},$$

the average output current is

$$I_{dc} = \omega C E_m/\pi,$$

and the per cent ripple is 121.2 per cent. The per cent ripple and short-circuit currents calculated from these expressions are in quite good agreement with the experimentally determined values.

Comparison of Half-Wave and Full-Wave Doublers

A comparison of the operating characteristics of the half-wave and full-wave[4] voltage doublers shows that throughout the normal operating range (ωCR greater than 10) the full-wave doubler offers a higher input power factor, lower maximum tube currents, slightly less ripple (and of higher frequency) in the output voltage, and slightly better voltage regulation; while the half-wave doubler offers lower peak inverse tube voltages, lower effective input currents, and a common input and output terminal allowing both the load and input source to be grounded if necessary.

Performance and Design

The performance of an assembled half-wave doubler may be predicted if the capacitance of the two equal condensers C_1 and C_2 and the load resistance R are known. Upon evaluation of the parameter (ωCR), the curves of Figs. 6 to 10 may be used to determine the operating characteristics of the circuit.

These curves may also be used to design a half-wave doubler to meet certain prescribed operating conditions. Often the input voltage and frequency, the output direct voltage, and the output direct current are specified. In addition, the application may restrict the per cent ripple allowable in the output voltage. Hence E_m, E_{dc}, and I_{dc} are specified together with a restriction on the per cent ripple r. From the curve of Fig. 7 the value of (ωCR) is fixed by the ratio (E_{dc}/E_m). The capacitance of the condensers may be found from (ωCR), the load resistance $R=(E_{dc}/I_{dc})$, and the supply frequency $\omega/2\pi$. The curves of Figs. 9 and 10 may be used to determine the peak inverse tube voltage and the maximum tube current, thus enabling the selection of rectifier tubes of proper inverse voltage and maximum current ratings. From Fig. 7 the per cent ripple in the output voltage may be found. If this value is greater than the per cent ripple allowable, the output voltage may be filtered, or some compromise in the specified current and voltages may be made so as to increase the value of (ωCR). The per cent ripple in the output can be materially reduced by placing a filter circuit between the output condenser and the load resistance. Insertion of a filter will cause some alteration in the operation of the doubler circuit as predicted from this analysis; however, the analysis will still afford a rather good approximation of the operating characteristics of the doubler.

Conclusions

On the basis of the experimental verifications obtained, the mathematical analysis is representative of the actual operation of the doubler circuit, and therefore, prediction of the performance of the doubler by the analysis seems justifiable. The assumption of no tube drop while conducting seems to have introduced little error, but for large tube drops an extension of this analysis would have to be used.

It has been shown that the design of this circuit and the predetermination of its performance is facilitated by the use of the results of this analysis.

Appendix I

Nomenclature

$\omega/2\pi$ = frequency of the alternating-voltage supply
t = time in seconds
$\theta = \omega t$
E_m = the maximum value of the alternating-voltage supply
e = the instantaneous value of the alternating-voltage supply
i_1 = the current flowing through tube T_1
i_2 = the current flowing through tube T_2
i = the input current, i_1+i_2
i_L = the current flowing through the load resistance R
i_c = the current through condenser C_2
e_c = the instantaneous voltage on condenser C_1
q_1 = the charge on condenser C_1
q_2 = the charge on condenser C_2
R = the load resistance in ohms
C = capacitance in farads of condensers C_1 or C_2
α = angle in radians at which the tube T_1 starts to conduct current
β = angle in radians at which the tube T_1 stops conducting
γ = angle in radians during which tube T_1 is conducting, $\beta-\alpha$
δ = angle in radians at which the tube T_2 starts to conduct current
ξ = angle in radians during which the tube T_2 is conducting, $-(\delta+\pi/2)$
$Z=\sqrt{4R^2+(1/\omega C)^2}$
$\lambda_1=\tan^{-1}(\omega CR)$, $0\leqq\lambda_1\leqq(\pi/2)$
$\lambda_2=\tan^1 (2\omega CR)$, $0\leqq\lambda_2\leqq(\pi/2)$
i_α = current i_L at angle α
i_β = current i_L at angle β
$e_{c\beta}$ = voltage e_c at angle β
e_1 = instantaneous inverse voltage on the tube T_1
e_2 = instantaneous inverse voltage on the tube T_2
e_{p1} = peak inverse voltage on tube T_1
e_{p2} = peak inverse voltage on tube T_2
θ_p = angle at which the peak inverse voltage occurs on the tube T_2
i_{m1} = the maximum current through tube T_1
i_{m2} = the maximum current through tube T_2
θ_m = angle at which the maximum current through tube T_1 occurs
E = the effective value of the alternating-voltage supply
I = effective value of the input current
E_{dc} = the average value of the load voltage
I_{dc} = the average value of the load current
E_0 = the effective value of the load voltage
I_0 = the effective value of the load current
E_{0a} = the effective value of the alternating component of the load voltage
I_{0a} = the effective value of the alternating component of the load current
r = the per cent ripple
P = the average input power

Appendix II

Analysis

(a). Angles at Which the Tubes Start and Stop Conducting

From angle α to angle β the equivalent circuit of Fig. 3(a) is in operation. The initial conditions at angle α are substituted in the solutions of the circuit equations to give

$$i_c/C_2 = \omega E_m \cos \omega t - i_1/C_1, \tag{1}$$

$$i_\alpha R = E_m \sin \alpha + E_m, \tag{2}$$

$$i_L = \frac{E_m}{Z}\cos(\theta-\lambda_2) + \left[i_\alpha - \frac{E_m}{Z}\cos(\alpha-\lambda_2)\right]\exp[-(\theta-\alpha)\cot\lambda_2]. \tag{3}$$

At angle β, $i_L=i_\beta$; substituting in (3),

$$i_\beta = \frac{E_m}{Z}\cos(\beta-\lambda_2) + \left[i_\alpha - \frac{E_m}{Z}\cos(\alpha-\lambda_2)\right]\exp[-(\beta-\alpha)\cot\lambda_2]. \tag{4}$$

From angle β to angle (α+360 degrees) the equivalent circuit of Fig. 3(b) is in operation. For this circuit, $i_L=i_\beta$ at angle β, and

$$i_L = i_\beta \exp[-(\theta-\beta)\cot\lambda_1]. \tag{5}$$

At angle β, $i_\beta=i_L=-i_c$, and by (1),

$$i_\beta = -\omega C E_m \cos\beta. \tag{6}$$

At angle (α+360 degrees), $i_L=i_\alpha$ because of the steady-state conditions imposed, hence

$$i_\alpha = i_\beta \exp\{[\beta-(\alpha+2\pi)]\cot\lambda_1\}. \tag{7}$$

Equations (2), (4), (6), and (7) form a system of equations in the unknowns i_α, i_β, α, and β, and by the substitution of $\gamma=\beta-\alpha$ and after some manipulation, may be reduced to

$$A_1 \sin\beta + B_1\cos\beta = 1 \tag{8}$$

$$A_2 \sin\beta + B_2\cos\beta = 1 \tag{9}$$

where

$$A_1 = -\frac{\sin^2\lambda_2}{2}\exp(\gamma\cot\lambda_2) + \left[\frac{\sin\lambda_2}{2}\sin(\gamma+\lambda_2)-\cos\gamma\right]$$

$$B_1 = -\left[\omega CR + \frac{\sin\lambda_2\cos\lambda_2}{2}\right]\exp(\gamma\cot\lambda_2) + \sin\gamma + \frac{\sin\lambda_2}{2}\cos(\gamma+\lambda_2)$$

$$A_2 = -\cos\gamma$$

$$B_2 = \sin\gamma - \omega CR\exp[-(2\pi-\gamma)\cot\lambda_1].$$

From the relations $\sin^2\beta+\cos^2\beta=1$, (8) and (9) may be combined to give

$$(A_1 - A_2)^2 + (B_2 - B_1)^2 - (A_1B_2 - A_2B_1)^2 = 0. \quad (10)$$

These coefficients A_1, A_2, B_1, and B_2 are functions of (ωCR) and the angle γ; hence γ is defined implicitly in (10) as a function of (ωCR). For assigned values of (ωCR), the corresponding values of γ may be determined by several methods.[7,8] For a specific value of (ωCR), γ may be found from (10), angle β from (8) and (9), and $\alpha=\beta-\gamma$. Therefore, angles α, β, and γ are functions only of (ωCR).

The angle δ at which tube T_2 starts to conduct is determined by use of the fact that e_c at angle δ is equal to e_c at angle β, and from (1) is

$$e_{c\beta} = E_m(\sin\beta + \omega CR\cos\beta) = E_m\sin\delta, \quad (11)$$

and

$$\delta = \sin^{-1}(\sin\beta + \omega CR\cos\beta). \quad (12)$$

Because tube T_2 is conducting from $\theta=\delta$ to $\theta=-\pi/2$, its interval of conduction ξ is

$$\xi = -(\pi/2 + \delta). \quad (13)$$

(b). Average Load Voltage

The average load current I_{dc} is

$$I_{dc} = \frac{1}{2\pi}\int_{\alpha}^{\alpha+2\pi} i_L d\theta. \quad (14)$$

From $\theta=\alpha$ to $\theta=\beta$ the load current is expressed by (3), and from $\theta=\beta$ to $\theta=\alpha+2\pi$ the load current is given by (5). With the substitution of these expressions, (14) becomes, upon integration,

$$I_{dc} = \frac{\omega CE_m}{2\pi}\left[1 + \frac{\sin(\beta+\lambda_1)}{\cos\lambda_1}\right], \quad (15)$$

and from (15) the ratio of the average load voltage to the maximum value of the alternating voltage (E_{dc}/E_m) is

$$E_{dc}/E_m = I_{dc}R/E_m = \frac{\omega CR}{2\pi}\left[1 + \frac{\sin(\beta+\lambda_1)}{\cos\lambda_1}\right]. \quad (16)$$

This ratio is a function only of (ωCR).

(c). Per Cent Ripple

The effective value E_{0a} of the alternating component of the output voltage may be found from the effective value of the output voltage E_0 and the average output voltage E_{dc} by use of the relation

$$E_{0a} = \sqrt{E_0^2 - E_{dc}^2}. \quad (17)$$

The per cent ripple is then

$$r = \frac{E_{0a}}{E_{dc}} \times 100 = \frac{\sqrt{E_0^2 - E_{dc}^2}}{E_{dc}} \times 100. \quad (18)$$

[7] D. L. Waidelich, "The numerical solution of equations," *Elec. Eng.*, vol. 60, pp. 480–481; October, 1941.
[8] I. T. Whitaker and G. Robinson, "The Calculus of Observations," Blackie and Son, Ltd., London, England, 1924, pp. 78–95.

The effective output voltage E_0 is evaluated by means of the effective output current I_0, which in turn may be evaluated from

$$I_0 = \sqrt{\frac{1}{2\pi}\int_{\alpha}^{\alpha+2\pi} i_L^2 d\theta}. \quad (19)$$

With the substitution of (3) and (5), (19) becomes upon integration

$$\begin{aligned} 2\pi\left(\frac{RI_0}{E_m}\right)^2 &= \tfrac{1}{4}\sin^2\lambda_2\left\{\frac{\gamma}{2}+\tfrac{1}{4}[\sin 2(\beta-\lambda_2)-\sin 2(\alpha-\lambda_2)]\right\} \\ &+\sin^2\lambda_2\{\cos\beta[\omega CR\cos\beta+\tfrac{1}{2}\sin\lambda_2\cos(\beta-\lambda_2)] \\ &+\cos\alpha[(\sin\alpha+1)-\tfrac{1}{2}\sin\lambda_2\cos(\alpha-\lambda_2)]\} \\ &+\omega CR\{-[\omega CR\cos\beta+\tfrac{1}{2}\sin\lambda_2\cos(\beta-\lambda_2)]^2 \\ &+[(\sin\alpha+1)-\tfrac{1}{2}\sin\lambda_2\cos(\alpha-\lambda_2)]^2\} \\ &-\frac{\omega CR}{2}(\sin\alpha+1)^2+\frac{(\omega CR)^3}{2}\cos^2\beta. \end{aligned} \quad (20)$$

The per cent ripple r can then be found from

$$r = \frac{\sqrt{(RI_0/E_m)^2 - (E_{dc}/E_m)^2}}{E_{dc}/E_m} \times 100 \quad (21)$$

where (E_{dc}/E_m) may be obtained from (16).

(d). Effective Input Current and Power Factor

The effective input current I is evaluated from

$$I = \sqrt{\frac{1}{2\pi}\int_{\delta}^{\delta+2\pi} i^2 d\theta}. \quad (22)$$

From angle α to angle β the input current i is i_1, which from Fig. 3(a) is

$$i_1 = i_L + R\frac{di_L}{dt}, \quad (23)$$

and by (3)

$$i_1=\omega CE_m\left\{\cos\theta+\frac{\sin\lambda_2}{2}\sin(\theta-\lambda_2) -\cot\lambda_2\left[\frac{\sin\lambda_2}{2}\cos(\alpha-\lambda_2)-\sin\alpha-1\right]\exp[-(\theta-\alpha)\cot\lambda_2]\right\}. \quad (24)$$

From angle β to angle δ the input current is zero. From δ until $\theta=-\pi/2$ the input current is i_2, which from Fig. 3(c) is

$$i_2 = \omega CE_m\cos\theta. \quad (25)$$

With the substitution of (24) and (25) into (22), the effective current may be found from (26).

The elements of the doubler circuit, with the exception of the load resistance, are assumed to take no average power. Therefore, the power input to the doubler is the same as that taken by the load resistance, which is

$$P = I_0^2R. \quad (27)$$

This product I_0^2R is evaluated in (20). The power input may also be found from

$$2\pi\left(\frac{RI}{\omega CRE_m}\right)^2 = -\frac{\pi}{4}-\frac{\delta}{2}-\frac{1}{4}\sin 2\delta+\frac{1}{2}[\gamma+\tfrac{1}{2}(\sin 2\beta-\sin 2\alpha)]$$
$$+\sin\lambda_2\cos\lambda_2\left[\frac{\sin^2\beta-\sin^2\alpha}{2}\right]-\frac{\sin^2\lambda_2}{2}[\gamma+\tfrac{1}{2}(\sin 2\beta-\sin 2\alpha)]$$
$$-2\sin\lambda_2\cos\lambda_2\left\{[-\cot\lambda_2\cos\beta+\sin\beta]\left[\omega CR\cos\beta+\frac{\sin\lambda_2}{2}\cos(\beta-\lambda_2)\right]\right.$$
$$\left.+\left[\frac{\sin\lambda_2}{2}\cos(\alpha-\lambda_2)-\sin\alpha-1\right][\cot\lambda_2\cos\alpha-\sin\alpha]\right\}$$
$$+\tfrac{1}{8}\sin^2\lambda_2\{\gamma+\tfrac{1}{2}[\sin 2(\alpha-\lambda_2)-\sin 2(\beta-\lambda_2)]\}$$
$$-\sin^2\lambda_2\cos\lambda_2\left\{[\tfrac{1}{2}\sin\lambda_2\cos(\alpha-\lambda_2)-\sin\alpha-1][\cot\lambda_2\sin(\alpha-\lambda_2)+\cos(\alpha-\lambda_2)]\right.$$
$$\left.-\left[\omega CR\cos\beta+\frac{\sin\lambda_2}{2}\cos(\beta-\lambda_2)\right][\cot\lambda_2\sin(\beta-\lambda_2)+\cos(\beta-\lambda_2)]\right\}$$
$$+\frac{\cot\lambda_2}{2}\left\{\left[\frac{\sin\lambda_2}{2}\cos(\alpha-\lambda_2)-\sin\alpha-1\right]^2-\left[\omega CR\cos\beta+\frac{\sin\lambda_2}{2}\cos(\beta-\lambda_2)\right]^2\right\}. \quad (26)$$

$$P=\frac{1}{2\pi}\int_{\delta}^{\delta+2\pi} ei\,d\theta. \quad (28)$$

The input power factor is, from (27),

$$\frac{I_0^2R}{EI}\times 100. \quad (29)$$

(e). Peak Inverse Tube Voltages

1. Tube T_1

The ratio of the peak inverse voltage on tube T_1 to the maximum value of the supply voltage (e_{p1}/E_m) is

$$\frac{e_{p1}}{E_m}=\omega CR\cos\beta\exp[(\beta-\delta-2\pi)\cot\lambda_1]. \quad (30)$$

2. Tube T_2

The ratio of the peak inverse voltage on tube T_2 to the maximum value of the supply voltage (e_{p2}/E_m) is given by

$$\frac{e_{p2}}{E_m}=\frac{\sin\lambda_2}{2}\cos(\theta_p-\lambda_2)$$
$$+\left[\sin\alpha+1-\frac{\sin\lambda_2}{2}\cos(\alpha-\lambda_2)\right]\exp[(-\theta_p-\alpha)\cot\lambda_2] \quad (31)$$

where θ_p is the angle at which the peak inverse voltage occurs and may be obtained from

$$0=\frac{\sin\lambda_2}{2}\sin(\theta_p-\lambda_2)$$
$$+\cot\lambda_2\left[\sin\alpha+1-\frac{\sin\lambda_2}{2}\cos(\alpha-\lambda_2)\right]$$
$$\exp[(-\theta_{p-\alpha})\cot\lambda_2]. \quad (32)$$

For values of (ωCR) greater than 10, angle θ_p is very nearly equal to β and a good approximation is given by

$$\frac{e_{p2}}{E_m}=-\omega CR\cos\beta. \quad (33)$$

(f). Maximum Tube Currents

1. Tube T_1

For values of (ωCR) greater than 13.2 the maximum current through tube T_1 is

$$i_{m1}=\omega CE_m\left\{\cos\alpha+\frac{\sin\lambda_2}{2}\sin(\alpha-\lambda_2)\right.$$
$$\left.-\cot\lambda_2\left[\frac{\sin\lambda_2}{2}\cos(\alpha-\lambda_2)-\sin\alpha-1\right]\right\}. \quad (34)$$

For (ωCR) less than 13.2 the maximum tube current is given by

$$i_1=\omega CE_m\left\{\cos\theta_m+\frac{\sin\lambda_2}{2}\sin(\theta_m-\lambda_2)\right.$$
$$-\cot\lambda_2\left[\frac{\sin\lambda_2}{2}\cos(\alpha-\lambda_2)-\sin\alpha-1\right]$$
$$\left.\exp[-(\theta_{m-\alpha})\cot\lambda_2]\right\}, \quad (35)$$

where θ_m is the angle at which i_1 is a maximum and may be found from

$$0=\sin\theta_m+\frac{\sin\lambda_2}{2}\cos(\theta_m-\lambda_2)$$
$$+\cot^2\lambda_2\left[\sin\alpha+1-\frac{\sin\lambda_2}{2}\cos(\alpha-\lambda_2)\right]$$
$$\exp[-(\theta_{m-\alpha})\cot\lambda_2]. \quad (36)$$

2. Tube T_2

For values of (ωCR) greater than 4.50 the maximum current through tube T_2 is

$$i_{m2}=-\omega CE_m\cos\delta. \quad (37)$$

For (ωCR) less than 4.50 the maximum tube current is given by

$$i_{m2}=-\omega CE_m. \quad (38)$$

Some Characteristics of a Stable Negative Resistance*

CLEDO BRUNETTI†, ASSOCIATE, I.R.E., AND LEIGHTON GREENOUGH‡, NONMEMBER, I.R.E.

Summary—*By employing positive feedback in a two-stage amplifier it is possible to obtain an input impedance which is equal to the impedance in the feedback circuit multiplied by a negative constant. For a resistance-capacitance feedback circuit the input impedance becomes a negative resistance in series with a negative capacitance. With the proper choice of circuit constants, the negative impedance can be made to approximate closely a pure negative resistance over any given frequency range. In the apparatus described, high stability is secured by the use of inverse feedback in addition to the positive-feedback loop.*

THERE is a need for a source of negative resistance that is independent of frequency and the normal variations in tubes and supply voltages.[1] Such a device would find application in the design of oscillators,[2] the improvement of the Q of tuned parallel circuits,[3] the measurement of circuit impedances at low and high frequencies,[4] the design of special networks,[5] and other uses.

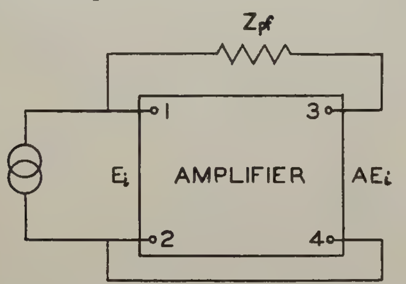

Fig. 1—Method of obtaining negative resistance.

The possibility of realizing such a negative resistance has already been pointed out in the literature.[3] If, in an ordinary amplifier, part of the output is coupled back to the input, it is possible to adjust the (positive) feedback and amplification so that the input impedance of the amplifier becomes a pure negative resisttance whose magnitude may be varied by changing either the positive feedback or the amplification. In fact, as pointed out by Crisson,[6] it is possible by this method to obtain an input impedance that is the negative counterpart of whatever impedance is used in the positive-feedback circuit. If the amplification is made substantially independent of tube and supply voltage variations by the employment of inverse feedback the resulting negative resistance will also be found to be independent of these quantities approximately to the same degree. It is the purpose of this paper to describe the characteristics of such a negative resistance and to point out both its advantages and limitations.

Consider the circuit of Fig. 1 where an impedance Z_{pf} is connected between the input and output terminals 1 and 3, respectively, of the two-stage amplifier. The equivalent circuit (assuming the impedance of the grid circuit of the first stage of the amplifier to be infinitely large) is shown in Fig. 2. If E_i is the input voltage and A the amplification, the resulting input current will be

$$I_i = \frac{E_i - AE_i}{Z_{pf}} \tag{1}$$

from which the input impedance becomes

$$Z_n = \frac{E_i}{I_i} = \frac{Z_{pf}}{1 - A}. \tag{2}$$

Thus, if A is real (i.e., has no phase shift) and is greater than unity, Z_n will be equal to Z_{pf} multiplied by a negative number. Z_n may be varied by changing either Z_{pf} or A, or both. In practice it is found best to vary both Z_{pf} and A, the former providing large-step variations in Z_n and the latter serving as a fine control. By keeping A low, optimum stability is obtained, provided, however, A does not approach unity, in which case slight variations in the magnitude of A would be magnified in Z_n.

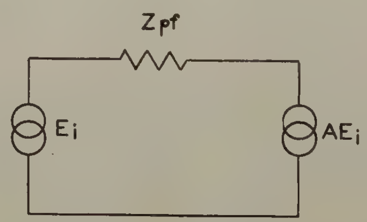

Fig. 2—Theoretical representation of amplifier as source of negative resistance.

Amplifier

The dependence of the magnitude of the negative impedance on the amplification may be found by differentiating (2) with respect to the real quantity A assuming Z_{pf} to remain constant. This yields

* Decimal classification: R132×R363.2. Original manuscript received by the Institute, April 15, 1942.

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[1] L. C. Verman, "Negative circuit constants," PROC. I.R.E., vol. 19, pp. 676–681; April, 1931.

[2] Cledo Brunetti, "The transitron oscillator," PROC. I.R.E., vol. 27, pp. 88–90; February, 1939.

[3] F. E. Terman, R. R. Buss, W. R. Hewlett, and F. C. Cahill, "Some applications of negative feedback with particular reference to laboratory equipment," PROC. I.R.E., vol. 27, pp. 649–655; October, 1939.

[4] H. Iinuma, "A method of measuring the radio-frequency resistance of an oscillatory circuit," PROC. I.R.E., vol. 18, pp. 537–543; March, 1930.

[5] S. P. Chakravarti, "The band-pass effect," *Wireless Eng.*, vol. 18, pp. 103–111; March, 1941.

[6] G. Crisson, "Negative impedances and the twin 21-type repeater," *Bell Sys. Tech. Jour.*, vol. 10, pp. 485–513; July, 1931.

$$\frac{dZ_n}{dA} = Z_n \frac{1}{1 - A} \qquad (3)$$

or

$$\frac{dZ_n}{Z_n} = \frac{dA}{1 - A} \cdot \qquad (4)$$

Equation (4) shows that, for A greater than unity, a given percentage variation in A will result in a larger percentage variation in Z_n. As A approaches unity the ratio of these variations increases without limit. This emphasizes the necessity of maintaining a constant amplification in the amplifier.

The circuit used in these measurements is shown in Fig. 3. It is strictly a conventional amplifier with simple positive and negative feedback applied to it as pointed out by Terman and others.[3] Amplification control is available by applying a variable amount of inverse feedback from the output back to the input and is accomplished with the variable resistor R_{nf}. This procedure yields good stability with respect to changes in supply voltages and frequency, as well as a low effective internal impedance in the output stage of the amplifier. A substantial amount of inverse feedback is employed so that low amplification and high stability result.

The amplification as a function of frequency for various values of R_{nf} is shown in Fig. 4. These measurements were made with a load impedance of 2500 ohms on the amplifier. In Fig. 5 is shown how the amplification may be controlled by varying R_{nf}. For an

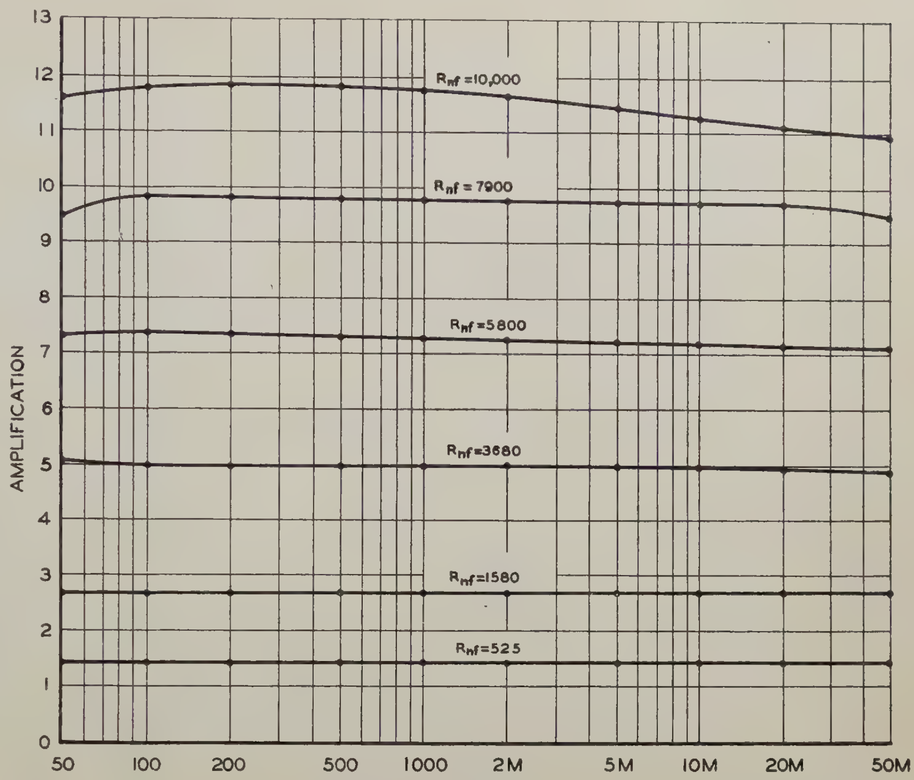

Fig. 4—Frequency response of amplifier for six different amplification settings. Load impedance is 2500 ohms.

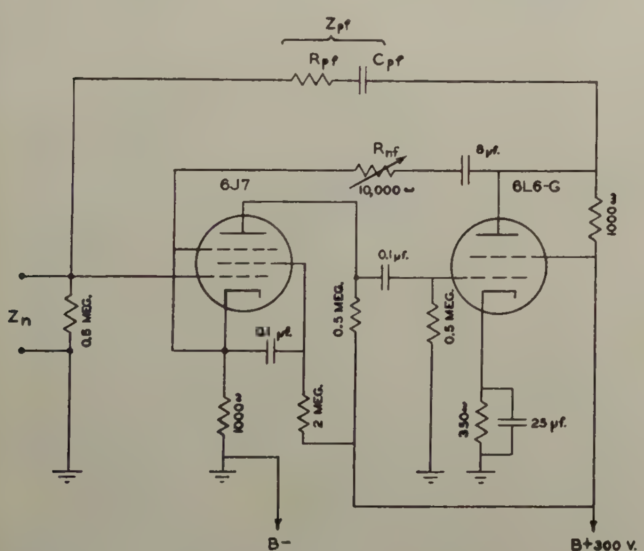

Fig. 3—Circuit for producing negative resistance.

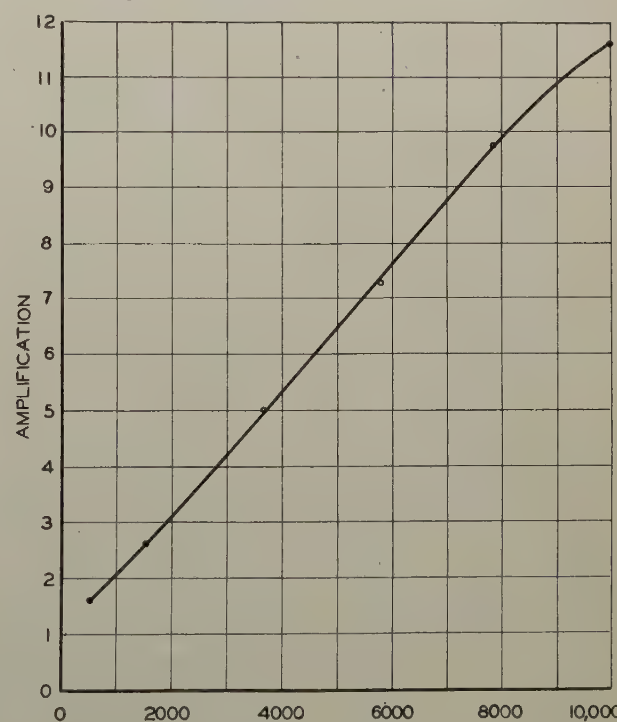

Fig. 5—Variation of amplification with inverse-feedback resistance R_{nf}.

amplification approaching a value of 10, the actual amplification variation is less than $2\frac{1}{2}$ per cent from 200 to 30,000 cycles per second. Below a value of 8 the variation is less than $2\frac{1}{2}$ per cent from 100 to 50,000 cycles.

Measurement of Negative Resistance

A simple circuit for obtaining the magnitude of negative resistance is to connect it in parallel with a

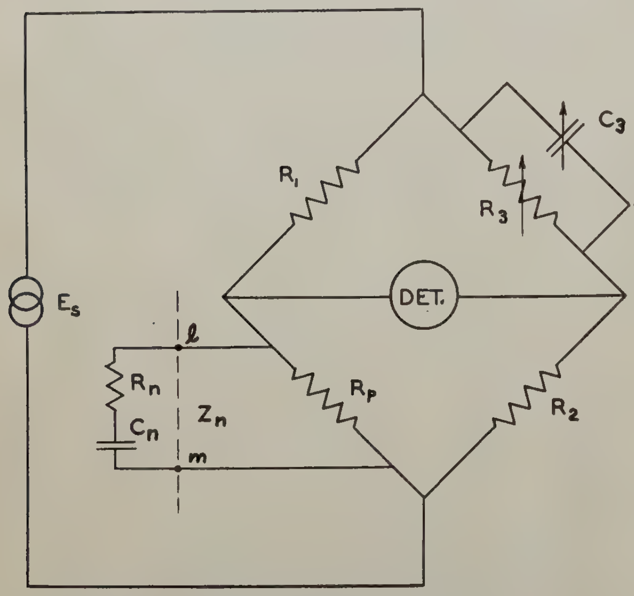

Fig. 6—Modified Maxwell bridge used for exact determination of Z_n. $R_1 = R_2 = 500$ ohms.

known positive resistance R_p and to measure the equivalent resistance of the combination on a Wheatstone bridge. From this R_n may easily be computed. R_n is the actual input resistance of the amplifier, and will have a negative sign. In this method, it is necessary that R_p be less in absolute value than R_n for otherwise natural oscillations will take place unless arrangements are made to keep the impedance of the external measuring circuit (which is in parallel with the combination) low.

This method of measurement does not take into account the phase shift resulting from the presence of the blocking condenser C_{pf}. Although usually made as large as possible, C_{pf} will still possess some reactance at low frequencies. A better circuit for precise determination of the complex impedance, and the one employed in these measurements, is the modified Maxwell bridge shown in Fig. 6. The negative impedance to be measured is connected to the left of terminals $l-m$ and is represented by R_n and C_n in series. If $Z_x = R_x + j\omega L_x$ is the equivalent series impedance of the parallel combination of R_n, C_n, and R_p, then, for balance, we have $R_x = R_1R_2/R_3$ and $L_x = R_1R_2C_3$. Using these quantities one may then compute the resistive and reactive components of the negative impedance Z_n from the equation

$$Z_n = -\frac{Z_xR_p}{Z_x - R_p}. \tag{5}$$

If Z_{pf} is a pure resistance of magnitude R_{pf}, and for zero phase shift in the amplifier, Z_n will be a pure negative resistance whose magnitude depends on both R_{pf} and the negative-feedback resistance R_{nf}. This dependence is shown in Fig. 7 where the negative impedance is plotted against R_{pf} for four values of R_{nf}. These curves show the wide range of negative impedance that may be obtained by this method simply by varying R_{pf}. The curvature at the lower portion of the curves is the result of the reactance of the blocking condenser and the variation of the amplification for low values of R_{pf}.

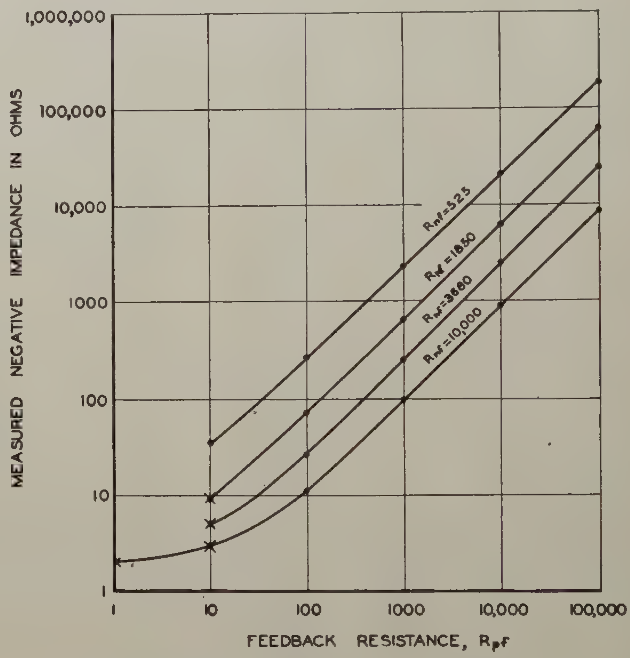

Fig. 7—Negative impedance as a function of positive-feedback resistance R_{pf} at 1000 cycles.

The ordinates of Fig. 7 represent the absolute magnitude of the negative impedance given by (2) which in general differs very little from the projection of the impedance on the negative real axis. The blocking condenser C_{pf}, as already noted, introduces some phase shift. However, with the exception of the points marked with a cross in Fig. 7, the capacitive reactance is negligible and the negative impedance and its negative-resistance component are for all practical purposes equal. For these points the impedance, measured at 1000 cycles, was accompanied by a phase shift of at most 5 to 15 degrees from a pure negative resistance. This phase shift may be reduced further if desired by increasing the size of the blocking condenser C_{pf}.

Some phase shift is also introduced in the amplifier at high frequencies by the shunting tube capacitance, and at the lower frequencies by the interstage coupling condenser.

The variation of negative resistance with frequency from 100 to 5000 cycles is shown in Fig. 8. Here also the absolute magnitude of the negative impedance is plotted for convenience. In the three upper curves for

which R_{pf} equals 1000, 10,000, and 100,000 ohms, respectively, the phase angle is less than 1 degree. For the curve for which R_{pf} equals 500 ohms the largest phase angle is four degrees at 100 cycles, while for the lower curve taken with R_{pf} equal to 50 ohms a maximum phase angle of 25 degrees occurs at 100 cycles.

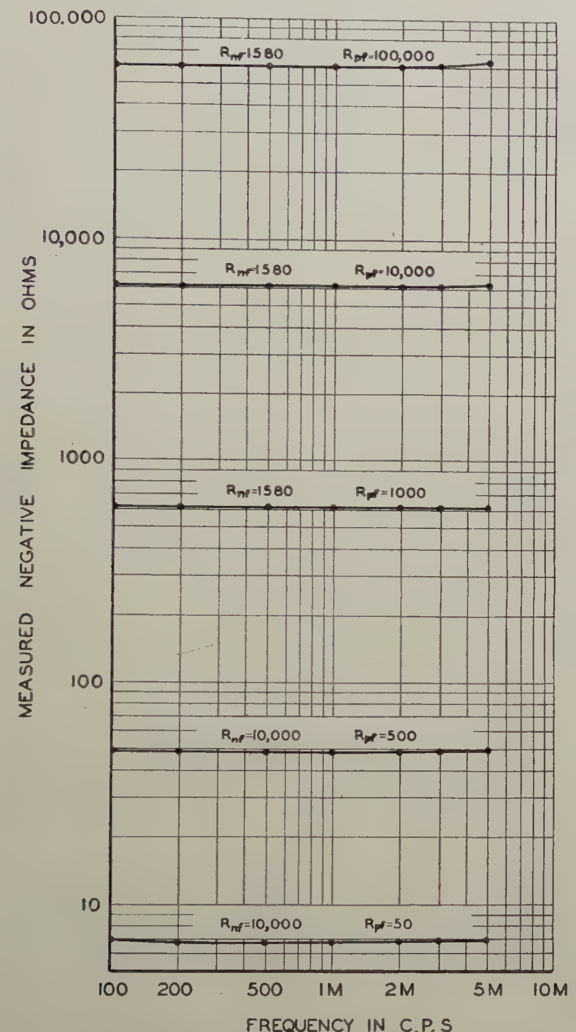

Fig. 8—Variation of negative impedance with frequency, for five representative cases.

The dependence of the negative resistance upon supply voltages is shown in Fig. 9. Here per cent changes are plotted rather than the actual value of negative resistance in order to emphasize the variations with supply voltages. These measurements were made at 1000 cycles. As is apparent, the large values of negative resistance are in general the most stable both in respect to variations in frequency as well as supply voltages.

Discussion

Low values of negative resistance may be obtained either by using low values of resistance R_{pf} in the positive-feedback circuit or by employing high amplification. Of these methods both result in unwanted phase changes produced by the blocking condenser C_{pf}. Raising the amplification is more desirable from the standpoint of keeping the phase angle of Z_n low as a higher value of R_{pf} may then be used to obtain a specified R_n. In this manner, the phase angle, being approximately proportional to the ratio of reactance in the positive-feedback circuit to the resistance in the same circuit, will be reduced. At low frequencies the reactance of C_{pf} may become appreciable. However, for most practical purposes where R_n is greater than ten ohms, the phase angle is negligible, even at a frequency of 100 cycles.

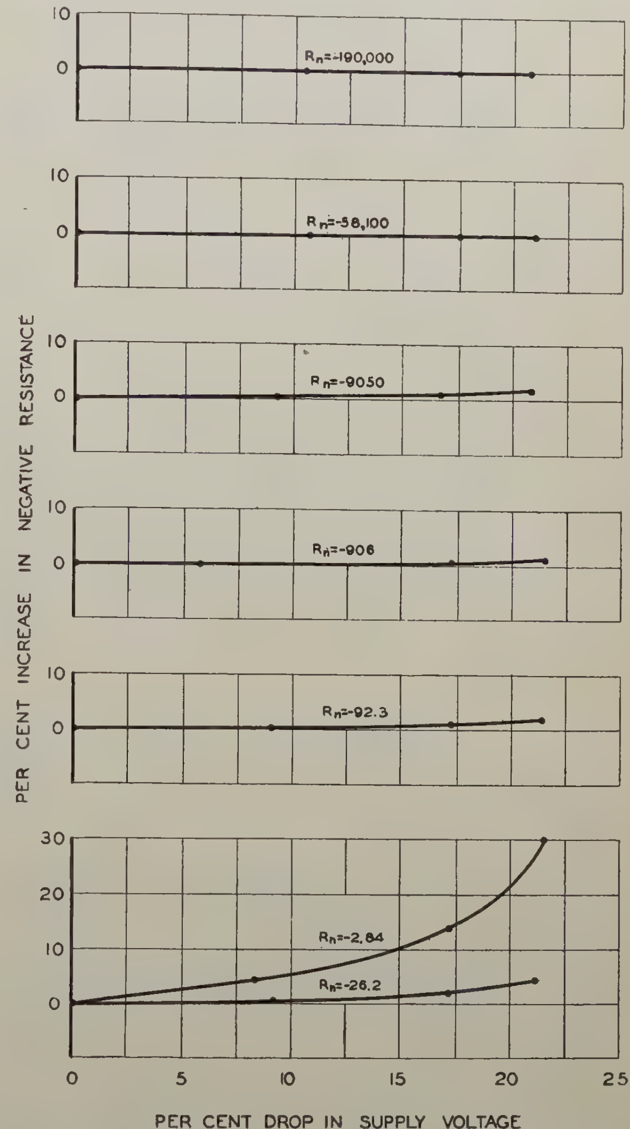

Fig. 9—Effect of supply-voltage variation on negative resistance.

A limitation in obtaining very low values of R_n is the effective internal impedance of the amplifier at its output terminals. This impedance (usually a resistance) must be included in Z_{pf}, as shown in Fig. 10. The use of sufficient inverse feedback will reduce Z_{int}, but the amplification is then also reduced. This makes it necessary to decrease R_{pf} to obtain the desired value of negative resistance. Z_{int} may vary from 5 to 20 ohms in the circuit of Fig. 3.

It should be kept in mind that R_{pf} effectively

constitutes the load on the amplifier. If a low value of R_n is obtained by using a low R_{pf}, the amplification will be somewhat less stable since less voltage is available at the output terminals for feedback, hence stabilization is not as good. Distortion, also, is greater.

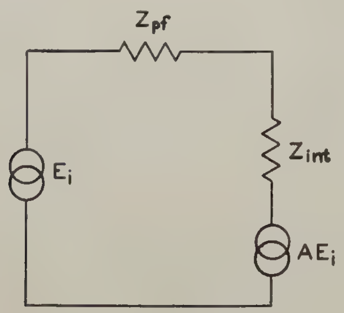

Fig. 10—Equivalent circuit including internal impedance of amplifier.

If a direct-current potential across the negative resistance is not objectionable, it is possible to modify the circuit so as to eliminate the phase angle introduced by C_{pf} as shown in Fig. 11. In this circuit the blocking condenser is placed in series with the high impedance of the grid circuit where it can do little harm.

As with all electrical elements, there is an upper limit to the alternating voltage which may be applied to the negative resistance when used as a circuit component. Because of the presence of the inverse feedback in the amplifier, alternating voltages of the order of 3 to 30 volts root-mean-square, depending upon the amplification used, may be applied to the

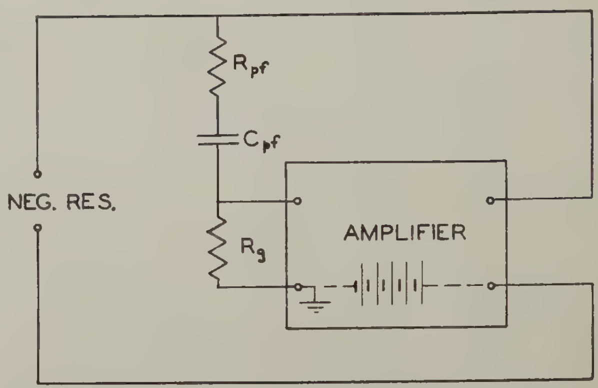

Fig. 11—Circuit to reduce effect of blocking-condenser reactance in series with R_{pf}.

device without introducing an appreciable amount of harmonic content in the output.

Summarizing, it can be stated that with this apparatus it is possible to obtain values of negative resistance from 1 ohm up to an unlimited value. The circuit, unlike the dynatron, is not dependent upon secondary emission for its operation. Stability with respect to frequency and supply voltages is good.

Thermal-Frequency-Drift Compensation*

T. R. W. BUSHBY†, ASSOCIATE, I.R.E.

Summary—*The conditions necessary for minimizing frequency drift with variation of ambient temperature are examined for various types of circuits. In fixed-tuned circuits the drift can be eliminated by a comparatively simple adjustment of the temperature coefficient of capacitance. In variable-tuned circuits, expressions for coefficient adjustment resulting in minimum integrated drift are given, together with simpler expressions resulting in an approximate minimum. It is shown to be practicable in some instances to design circuits in which the drift is better than ten parts per million per degree centigrade, over normally used frequency ranges, by complementary adjustment of the temperature coefficients of inductance and capacitance. Frequency drift in superheterodyne receiver circuits is discussed, and it is found that the local-oscillator circuit is peculiarly adaptable to drift correction by reason of the complex nature of its capacitance network. Such padded circuits can be very effectively corrected by simple adjustment of the various capacitive coefficients, the drift factor being tracked in a manner analogous to the frequency tracking. The padding of variable-capacitance-tuned circuits for the express purpose of drift correction presents the simplest means of minimizing thermal drift in such circuits. The necessary expressions are given, together with the results of some experimental work.*

I. Introduction

In THE parallel resonant circuit of Fig. 1, the frequency stability is dependent, among other things, on the temperature of the various circuit components.

For instance, when such a circuit is used to control the frequency of a valve generator, during the first few minutes after switching on the valve, there is a frequency change caused mainly by capacitance increase, which is due to the expansion of the valve electrodes.

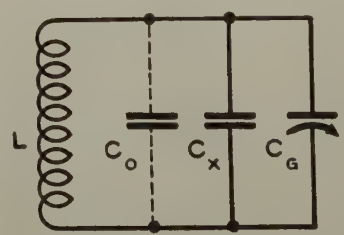

Fig. 1—Variable-capacitance-tuned circuit having thermal-drift-compensating condenser C_x.

Methods for reducing such drift have already been discussed[1,2] and it is not the purpose of this paper to deal further with it.

After a sufficient lapse of time at a constant ambient temperature, the frequency becomes stable, provided that other factors introduce no drift. Subsequently, changes in ambient temperature will cause changes in

* Decimal classification: R381×536. Original manuscript received by the Institute, April 7, 1942.

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[1] M. L. Levy, *Electronics*, vol. 12, p. 15; 1939.

[2] Gramophone Company, *Electronics and Television and Short-Wave World*, vol. 13, p. 407; 1940.

circuit inductance and capacitance, resulting in further frequency drift.

It is usual to express the temperature coefficients of capacitance and inductance as a number of parts per million per degree centigrade, referred to a temperature of 20 degrees centigrade, so that a coil having a temperature coefficient of +100 increases its inductance by 100 parts in a million, when raised in temperature from 20 to 21 degrees centigrade.

Ordinarily, the reference point may be arbitrarily selected with negligible error, but for extremes of temperature and/or large coefficients, the reference temperature of 20 degrees centigrade should be adhered to.

In Fig. 1, let $\alpha=(dC/dT)(10^6/C)$ be the numerical value of the temperature coefficient of the total capacitance C (where $C=C_0+C_g+C_x$), and $\beta=(dL/dT)(10^6/L)$, the corresponding coefficient for the inductance.

The resonant frequency at the reference temperature is $f=1/2\pi\sqrt{LC}$, from which it may be shown that the frequency drift due to changing temperature is given by

$$\theta = -0.5(\alpha+\beta) \tag{1}$$

where $\theta=(df/dT)(10^6/f)$, the temperature coefficient of frequency, in parts per million per degree centigrade.

II. Fixed-Tuned Circuits

In a fixed-tuned circuit $\theta=0$ when $\alpha+\beta=0$, and the resonant frequency will then be independent of temperature.

Inductances have, in general, a positive coefficient, so that the total capacitance must usually have a negative coefficient of the same numerical value, in order that there may be no frequency drift with temperature.

The availability of ceramic materials having dielectric constants which vary inversely with temperature, has made possible the manufacture of condensers having negative temperature coefficients. The capacitance of such condensers decreases with increasing temperature, and in this respect they differ from almost all other circuit components.

As it is unlikely that a condenser having the exact numerical value of coefficient will be available, it will be necessary to use a parallel combination of two condensers, one of positive coefficient, and one of negative coefficient, to obtain the required result. Therefore, let $C=C_1+C_2\cdots+C_n$, and let $\alpha_1, \alpha_2\cdots, \alpha_n$ be the respective temperature coefficients. Then α_c, the temperature coefficient of C, is given by

$$\alpha_C = \frac{\alpha_1\cdot C_1+\alpha_2\cdot C_2\cdots+\alpha_n\cdot C_n}{C}. \tag{2}$$

To take a fairly general practical case, let

C_0 = self-capacitance of the inductance
C_x = drift-compensating condenser
C_p = portion of C_x having α positive
C_n = portion of C_x having α negative
C_t = capacitance of the trimmer condenser
C_s = stray capacitance
α_0, α_p etc. = respective temperature coefficients
β = temperature coefficient of inductance
θ = temperature coefficient of frequency

All temperature coefficients are in parts per million per degree centigrade, and all capacitances are in micromicrofarads.

As the self-capacitance is inseparable from the inductance, it is not possible to measure either β or α_0 separately. The observed value β' for the temperature coefficient of inductance is

$$\beta' = \beta + \frac{\alpha_0\cdot C_0}{C} \tag{3}$$

where C is the total capacitance of the test circuit.

Since β varies with frequency, it must be measured at the frequency at which the inductance is to be used; therefore the test circuit will necessarily have the same total capacitance as the circuit undergoing analysis. Then θ will be zero when

$$\beta' = -\left[\frac{\alpha_p\cdot C_p+\alpha_n\cdot C_n+\alpha_t\cdot C_t+\alpha_s\cdot C_s}{C}\right] \tag{4}$$

which is equivalent to saying that α_0 automatically becomes part of β', and can be disregarded, provided that C_0 is not omitted from the total circuit capacitance C.

The apparatus used for the measurement of individual component coefficients is substantially that described by Leonard.[3] The component under test is placed in a small temperature chamber and connected to a two-terminal oscillator of good short-period stability.

This oscillator is initially set to beat with a substandard crystal-controlled oscillator, and the beatnote observed with changing temperature of the component. An ordinary receiver, cathode-ray oscilloscope, and beat-frequency oscillator is used to measure the beat frequency.

Then, for an inductance, $\beta=-2\theta$, and for a capacitance, $\alpha=-2C\theta/C_c$, where C_c is the capacitance of the condenser under test, and C is the total circuit capacitance of the test oscillator.

Complete assemblies are checked for thermal drift by being placed in a larger temperature chamber, and loosely coupled to the receiver. With the latter at some feet distant, deliberate coupling is usually unnecessary.

As will be seen throughout this paper, a knowledge of individual coefficients is not, in general, necessary for

[3] S. C. Leonard, *Electronics*, vol. 11, p. 18; 1938.

drift compensation, but it is useful for checking purposes and for *ab initio* design.

All components except C_s can be dissociated from the circuit for individual measurement, but the value of α_s can only be determined indirectly. The method used is to place in the circuit temporarily a condenser C_x of known temperature coefficient α_x and check the thermal drift for the complete circuit. α_s is then determined from (5).

$$\alpha_s = -\left[\frac{C(2\theta+\beta')+\alpha_x\cdot C_x+\alpha_t\cdot C_t}{C_s}\right]. \tag{5}$$

Then for $\theta=0$

$$\alpha_{x'} = -\left(\frac{C\beta'+\alpha_s\cdot C_s+\alpha_t\cdot C_t}{C_x}\right) \tag{6}$$

where $\alpha_{x'}$ is the value of α_x for $\theta=0$. Combining (5) and (6) to eliminate α_s gives

$$\alpha_{x'} = \frac{2C\cdot\theta}{C_x}+\alpha_x \tag{7}$$

and from (7) and (2)

$$C_p = \frac{2C\cdot\theta+C_x(\alpha_{x'}-\alpha_n)}{\alpha_p-\alpha_n}. \tag{8}$$

Only two heat runs are necessary, one for the complete circuit and one for C_x.

Therefore, in fixed-tuned circuits, thermal frequency drift may be eliminated by a simple adjustment of the temperature coefficient of a portion of the capacitance, subject to a realizable value for $\alpha_{x'}$. When the figure found for $\alpha_{x'}$ is not realizable, some circuit change must be made, either in electrical values or in temperature coefficients.

III. Variable-Capacitance-Tuned Circuits

When the circuit has to be designed to cover a range of frequencies, exact compensation at all points within the range is not possible, since the temperature coefficient of a variable condenser varies with setting, while the temperature coefficient of inductance alters with frequency.

The following notation applies to the circuit of Fig. 1.

- f_L = lowest resonant frequency
- f_H = highest resonant frequency
- θ = temperature coefficient of frequency
- $\theta_L = \theta$ at f_L in a test heat run of the complete circuit
- $\theta_{L'} = \theta$ at f_L after circuit correction
- $\theta_H = \theta$ at f_H in a test heat run of the complete circuit
- $\theta_{H'} = \theta$ at f_H after circuit correction
- C_L = total circuit capacitance at f_L
- C_H = total circuit capacitance at f_H
- C_x = capacitance of the drift-compensating condenser
- α_x = temperature coefficient of C_x in the test heat runs
- $\alpha_{x'}$ = temperature coefficient of C_x after circuit correction
- C_{max} = capacitance of C_g at f_L
- α_{max} = temperature coefficient of C_{max}
- C_{min} = capacitance of C_g at f_H
- α_{min} = temperature coefficient of C_{min}
- β_L = temperature coefficient of inductance at f_L
- β_H = temperature coefficient of inductance at f_H
- $k = C_L/C_H$

All temperature coefficients are in parts per million per degree centigrade, and all capacitances are in micromicrofarads.

If we make the simplifying assumption that θ varies linearly with frequency, it can be shown that it will have its minimum value, integrated throughout the range, when $\theta_{L'}^2+\theta_{H'}^2$ is a minimum, and that $\theta_{L'}$ and $\theta_{H'}$ will then have opposite signs. That is, the drift will be in opposite directions at the frequency extremes, and will therefore be zero at some frequency within the range. The assumption of linearity has been found to be sufficiently true for all practical purposes, except when the integrated drift is very low.

From (1) and (2),

$$\theta_{L'}^2+\theta_{H'}^2 = \left(\beta_L+\frac{\alpha_{x'}\cdot C_x+\alpha_{max}\cdot C_{max}}{C_L}\right)^2 + \left(\beta_H+\frac{\alpha_{x'}\cdot C_x+\alpha_{min}\cdot C_{min}}{C_H}\right)^2 \tag{9}$$

which has its minimum value when

$$\alpha_{x'} = \frac{2C_L(\theta_L+k\cdot\theta_H)}{C_x(k^2+1)}+\alpha_x \tag{10}$$

which reduces to (7) for the fixed-tuned circuit. When $\alpha_{x'}$ has its optimum value, then

$$\theta_{L'} = \frac{k^2\cdot\theta_L-k\cdot\theta_H}{k^2+1} \tag{10a}$$

and

$$\theta_{H'} = \frac{\theta_H-k\cdot\theta_L}{k^2+1}. \tag{10b}$$

$\theta_{H'}$ will usually have a small positive value and $\theta_{L'}$ a larger negative one, which is a desirable condition. Therefore, a close approximation to optimum drift is had by putting $\theta_{H'}=0$ either experimentally, or by (10c), where

$$\alpha_{x'} = \frac{2C_H\cdot\theta_H}{C_x}+\alpha_x \tag{10c}$$

which is the same expression as (7). When $\alpha_{x'}$ has this value, then

$$\theta_{L'} = \theta_L - \theta_H/k. \tag{10d}$$

It will be noted that the terms of the expressions are such that the correction possibilities can be fully explored after the initial experimental work is done. The latter consists only of three heat runs to determine θ_L, θ_H, and α_x, and three capacitance measurements for C_L, C_H, and C_x.

The drift at other frequencies within the range will usually be less than at the extremes, depending upon the linearity of the curve. The deviation may amount to about ten parts per million, so that when the corrected drift is of this order, the assumption of linearity is no longer justified, and the final correction is best made experimentally. This necessitates taking some heat runs at intermediate frequencies, unless the individual coefficients are known for such frequencies.

In the latter case, the drift throughout the range can be determined graphically, by plotting the effect of each circuit component individually and summing them as shown in Fig. 2, which represents a practical example where the approximate values are as in Table I.

TABLE I

$C_L = 360\mu\mu f$	$C_{max} = 310\mu\mu f$	$\beta_L = +20$
$C_H = 90\mu\mu f$	$\alpha_{max} = +130$	$\beta_H = +35$
$C_x = 50\mu\mu f$	$C_{min} = 40\mu\mu f$	$\theta_L = -70$
$\alpha_x = +40$	$\alpha_{min} = +250$	$\theta_H = -85$
α_x' is found to be -300, when θ_L' becomes -45 and θ_H' $+10$.		

In the figure, curve A shows the effect on the frequency drift of the sum of the coefficients of the various capacitances, plotted against resonant frequency. The individual curves are omitted in the interests of clarity. Curve B is a similar plot for the capacitance, after the inclusion of the required negative-coefficient condenser for C_x, and curve C is the final curve, including the coil coefficient, being the algebraic sum of -0.5α and -0.5β.

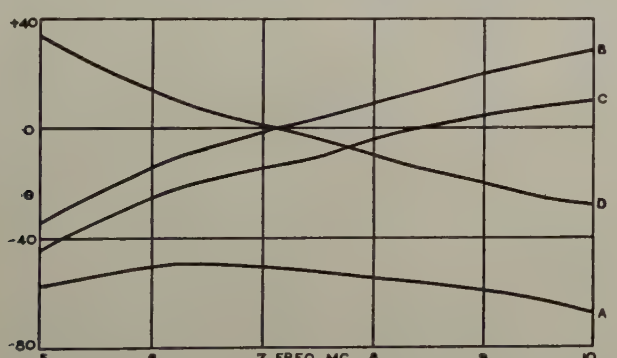

Fig. 2—Illustrating a practical example of optimum drift correction in a circuit using stock broadcast receiver parts.

From the final curve it is seen that $\theta_{L'}$ is -45 and $\theta_{H'}$ is $+10$, as already found, and in addition, the drift at intermediate frequencies is readily observed. The circuit used for Fig. 2 is fairly typical of those encountered when using average grade components designed for commercial receiver use.

Thus a considerable reduction in drift, at all frequencies, is possible by a comparatively simple adjustment of the temperature coefficient of a portion of the fixed capacitance. No further improvement can be had unless some change is made in the temperature coefficient of inductance.

IV. Low-Drift Variable-Tuned Circuits

Referring again to Fig. 2, in which curve B represents the drift due to the capacitance coefficient, it is seen that a coil having a coefficient such that -0.5β is represented by curve D, will result in zero drift at all frequencies.

Zero drift in variable-tuned circuits therefore depends entirely on the practical possibilities of so arranging the individual coefficients, that the algebraic sum is zero for all required frequencies. A coil is required having coefficients at the frequency extremes which are related by

$$\beta_L = \frac{C_H \cdot \beta_H - (\alpha_{max} \cdot C_{max} - \alpha_{min} \cdot C_{min})}{C_L} \tag{11}$$

which is derived from (9) when $\theta_{L'} = \theta_{H'} = 0$.

In some cases, the design of inductances which will meet the requirements to a sufficient degree of accuracy, is not unduly difficult, since it is fortunate that, in general, β increases with increasing frequency, while α can be made to decrease with increasing frequency, by adjustment of the temperature coefficient of a portion of the capacitance.

A practical example (Figs. 1 and 3) will illustrate this. The circuit constants were approximately as in Table II.

TABLE II

$C_L = 192\mu\mu f$	$\alpha_x = 0$	$C_{min} = 36\mu\mu f$
$C_H = 48\mu\mu f$	$C_{max} = 180\mu\mu f$	$\alpha_{min} = +165$
$C_x = 12\mu\mu f$	$\alpha_{max} = +65$	

Fig. 3 is constructed in the same manner as Fig. 2. A is the drift attributable to the total capacitance before correction, and B the drift due to capacitance after correction. In this case, the correction is arbitrary, as the coil design is not considered until the capacitance drift has been plotted. A number of B curves can be laid out, according to the available choice of

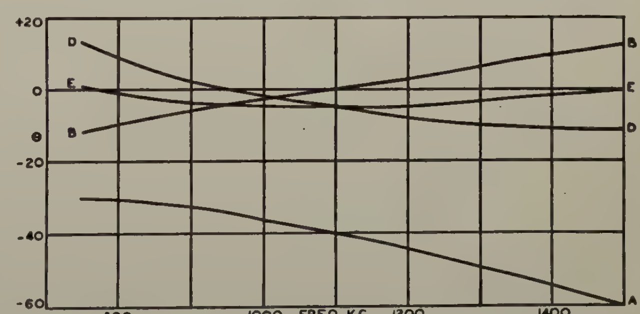

Fig. 3—Illustrating a practical example of drift correction by complementary adjustment of the temperature coefficients of inductance and fixed capacitance.

temperature coefficients for C_x. Only the one finally chosen is shown. It corresponds to $\alpha_{x'} = -600$.

The mirror images of the B curves about the zero θ axis will then provide design data for a number of possible coils, any one of which can be used, with its appropriate value for $\alpha_{x'}$, to provide a variable-frequency zero-drift circuit. Curve D gives the characteristic of the chosen coil, and E shows the drift for the whole circuit throughout the range.

The latter curve is plotted from the known coefficients of the various components. Experimental checks showed the drift throughout the range to be very small.

The inductance in this instance consisted of four litz pies on a bakelized-paper former, and it had zero temperature coefficient at about 950 kilocycles. The negative coefficient at the lower frequencies is ascribed mainly to the axial expansion of the former, whereby the pies are moved further apart with increasing temperature.[4] Other components of the coefficient become increasingly significant at higher frequencies, and the effect of the expansion of the former is more than offset by them.[5]

For the example given, it is obvious that the drift throughout the tuning range will be due entirely to practical considerations of coefficient tolerances in the various components.

The necessity for adjusting the temperature coefficient of inductance may be avoided by padding the circuit, as is done in the case of the superheterodyne local-oscillator. These possibilities are explored in Section VII, which deals with the latter circuits.

V. Variable-Inductance-Tuned Circuits

Although (10) is arrived at on the basis of a variable-capacitance circuit, it is seen to be general and applies equally well to the case of the variable-inductance-tuned circuit.

Where the method of varying the inductance does not alter the self-capacitance of the inductance, or the stray capacitance, (10) reduces to

$$\alpha_{x'} = \frac{C(\theta_L + \theta_H)}{C_x} + \alpha_x \tag{12}$$

whence

$$\theta_{L'} = -0.5(\theta_H - \theta_L) \tag{12a}$$

and

$$\theta_{H'} = 0.5(\theta_H - \theta_L). \tag{12b}$$

Usually C_0 (and also possibly C_s) does vary with inductance variation, but if the change is small in relation to the total capacitance, the error involved in using (12) instead of (10) will also be small.

The correction procedure of Section IV cannot be applied to the variable-inductance-tuned circuit, since α is constant throughout the range; therefore, for $\theta = 0$ at all frequencies, β must also be constant with varying frequency, which is not possible. The best correction for such circuits is obtained when (10) or (12) is applied, after the variation of β is reduced to a minimum. This is seen from (11), which becomes $\beta_L = \beta_H$.

A method exists for rendering such circuits effectively driftfree, by a combination of capacitances of different coefficients, connected to different points on the inductance.[6]

Variable-inductance tuning has a very useful application to circuits which are required to have a minimum of drift with "warm-up" period. The total capacitance in the circuit can be fixed at a sufficiently high value to reduce the short-period drift to negligible proportions, avoiding the necessity for special measures to overcome such drift.[7]

VI. The Superheterodyne Receiver

A superheterodyne receiver is correctly aligned when the frequency of the local oscillator is equal to the sum of the intermediate and signal frequencies, provided that the usual practice of having the oscillator frequency higher than the signal frequency is employed.

Changes in one or more of the involved frequencies will result in attenuation or loss of the signal. Assuming that the signal frequency remains constant, and disregarding the possibilities of automatic frequency control of the local oscillator, circuit drift of the various receiver channels will be separately considered.

Drift in the intermediate-frequency channel results in reduction of the signal amplitude subsequent to conversion, the attenuation being dependent on the amount of drift, and the selectivity of the channel. Being fixed-tuned circuits, they are readily corrected by the method given in Section II.

Designing the intermediate-frequency channel so as to pass a band of frequencies without attenuation, permits some drift in either the signal or oscillator frequencies, or both, without effect on the signal strength, and is a useful means of allowing for the residual drift, after other means of circuit correction have been exhausted.

Drift in the signal-frequency circuits attenuates the incoming signal prior to conversion. As in the case of the intermediate-frequency circuits, the degree of attenuation is dependent on the amount of drift, and the circuit selectivity. Though often neglected in practice, the effect can be quite serious with highly selective preconverter circuits. Methods for reducing this drift have also been discussed.

VII. Padded Circuits

The local-oscillator circuit deserves special attention, since the capacitance network is complicated by the

[4] W. H. F. Griffiths, *Wireless Eng. and Exp. Wireless*, vol. 11, p. 305; 1934.
[5] Tj. Douma, *Philips Transmitting News*, vol. 5, p. 20; 1938.
[6] Australian Patent Application No. 3070/41.
[7] P. Ware, *Electronics*, vol. 10, 12, p. 22; 1937.

necessity for frequency tracking with the signal-frequency circuits, the circuit being usually as shown in Fig. 4, to which circuit the following notation applies:

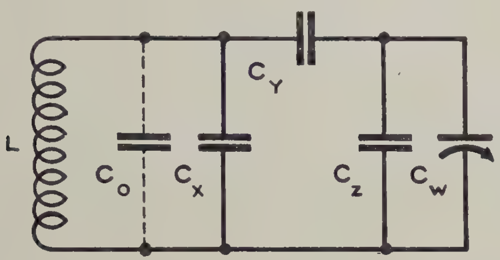

Fig. 4—Padded circuit as used in the superheterodyne local oscillator. C_y and C_x or C_y and C_z may be temperature compensated to give the circuit a very low drift factor over a range of frequencies.

f_L = lowest resonant frequency
f_H = highest resonant frequency
θ = temperature coefficient of frequency
$\theta_L = \theta$ at f_L in a test heat run of the complete circuit
$\theta_{L'} = \theta_L$ after circuit correction
$\theta_H = \theta$ at f_H in a test heat run of the complete circuit
$\theta_{H'} = \theta_H$ after circuit correction
β = temperature coefficient of inductance
C_0 = self-capacitance of the inductance L
$C_g = C_w + C_z$
C_{max} = capacitance of C_g at f_L
C_{min} = capacitance of C_g at f_H

$$C_L = C_0 + C_x + \frac{C_y \cdot C_{max}}{C_y + C_{max}}$$

$$C_H = C_0 + C_x + \frac{C_y \cdot C_{min}}{C_y + C_{min}}$$

α_x, α_y, etc. = respective temperature coefficients during test heat runs
$\alpha_{x'}$ $\alpha_{y'}$, etc. = respective temperature coefficients after circuit correction

$$a = \frac{C_{max}^2}{(C_y + C_{max})^2} \qquad b = \frac{C_{min}^2}{(C_y + C_{min})^2}$$

$$k = C_L/C_H = (f_H/f_L)^2 \qquad p = a/b$$

$$m = \frac{C_y^2}{(C_y + C_{max})^2} \qquad n = \frac{C_y^2}{(C_y + C_{min})^2}$$

All temperature coefficients are in parts per million per degree centigrade, and all capacitances are in micromicrofarads.

The coefficient α_C of a series combination of two condensers, C_y having coefficient α_y, and C_g having coefficient α_g can be shown to be

$$\alpha_C = \frac{\alpha_y \cdot C_g + \alpha_g \cdot C_y}{C_y + C_g} \qquad (13)$$

Therefore, θ for this circuit is given by

$$\theta = -0.5\left[\beta + \frac{C_x \cdot \alpha_x}{C} + \frac{C_g^2 \cdot C_y \cdot \alpha_y + C_y^2(\alpha_w \cdot C_w + \alpha_z \cdot C_z)}{C(C_g + C_y)^2}\right]. \qquad (14)$$

The determination of θ is possible when all of the individual coefficients are known, but as in the previous case, the coefficient of the stray capacitance cannot be measured directly, therefore experimental heat runs of the complete circuit are essential.

This circuit opens up some very interesting possibilities for temperature compensation. In Fig. 1, there are only three possible variables, the temperature coefficients of inductance and of variable and fixed capacitance. The adjustment of the coefficients of the inductance or the variable capacitance is a comparatively difficult matter, and if these are eliminated from consideration, only C_x remains, so that (10a) and (10b) give the optimum correction figures.

But in Fig. 4, different effects are had according to which particular coefficient is adjusted. The possible variables are now five in number, and they can be operated on either singly, or in any convenient combination.

Eliminating the coefficients of the inductance and variable capacitance, as in the previous instance, still leaves the three condensers C_x, C_y, and C_z available for correction purposes. Considering first the use of these condensers individually, the necessary information for C_x alone is given by (10). Expressions (15) and (16), respectively, have been developed using C_y and C_z alone.

Equation (15) is obtained by setting up expressions for $\theta_{L'}$ and $\theta_{H'}$ from (14), summing their squares, and finding a value for $\alpha_{y'}$ which makes the sum a minimum. This gives

$$\alpha_{y'} = \frac{2C_L(a \cdot \theta_L + k \cdot b \cdot \theta_H)}{C_y[a^2 + (kb)^2]} + \alpha_y. \qquad (15)$$

When $\alpha_{y'}$ has its optimum value, then

$$\theta_{L'} = \frac{k^2 \cdot \theta_L - k \cdot p \cdot \theta_H}{p^2 + k^2} \qquad (15a)$$

and

$$\theta_{H'} = \frac{p^2 \cdot \theta_H - k \cdot p \cdot \theta_L}{p^2 + k^2}. \qquad (15b)$$

The effect of the padder on the temperature coefficient is usually such that $\theta_{L'}$ has a small positive value and $\theta_{H'}$ a larger negative value, so that it is not as suitable as C_x or C_z for compensation purposes by itself. A close approximation to minimum integrated drift results when $\theta_{L'}$ is made zero, either experimentally or by means of (15c) where

$$\alpha_{y'} = \frac{2C_L \cdot \theta_L}{a \cdot C_y} + \alpha_y \qquad (15c)$$

then

$$\theta_{H'} = \theta_H - k\cdot\theta_L/p. \tag{15d}$$

Equation (16) is obtained similarly to (15) and gives

$$\alpha_{z'} = \frac{2C_L(m\cdot\theta_L + k\cdot n\cdot\theta_H)}{C_z[m^2 + (kn)^2]} + \alpha_z \tag{16}$$

when

$$\theta_{L'} = \frac{(kn)^2\theta_L - k\cdot m\cdot n\cdot\theta_H}{m^2 + (k_n)^2} \tag{16a}$$

and

$$\theta_{H'} = \frac{m^2\cdot\theta_H - k\cdot m\cdot n\cdot\theta_L}{m^2 + (kn)^2}. \tag{16b}$$

As in the case of the adjustment of $\alpha_{x'}$, a close approximation to minimum integrated drift results from putting $\theta_{H'} = 0$, when

$$\alpha_{z'} = \frac{2C_H\cdot\theta_H}{n\cdot C_z} + \alpha_z \tag{16c}$$

and

$$\theta_{L'} = \theta_L - m\cdot\theta_H/k\cdot n. \tag{16d}$$

VIII. Temperature-Coefficient Tracking

Either C_x, C_y, or C_z can be used to obtain some measure of drift reduction, but a combination of C_x and C_y (expression (17)), or C_z and C_y (expression (18)) can be so adjusted as to give a very low coefficient throughout a range of frequencies. Some work of this nature has already been carried out in a specific instance.[8]

Equations (17) and (17a) are obtained by equating the expressions for $\theta_{L'}$ and $\theta_{H'}$ to zero, when

$$\alpha_{y'} = \frac{2C_H(k\cdot\theta_L - \theta_H)}{C_y(a - b)} + \alpha_y \tag{17}$$

and

$$\alpha_{x'} = \frac{2C_H(p\cdot\theta_H - k\cdot\theta_L)}{C_x(p - 1)} + \alpha_x. \tag{17a}$$

Subject to realizable values for $\alpha_{y'}$ and $\alpha_{x'}$, the drift can be zero at both extremes of the range, and will depart from zero only in so far as the assumption of linearity of θ versus frequency is unjustified. In practice, deviations of about ten parts per million may be expected. If the departure is too great, putting $\theta = 0$ at two suitable frequencies within the range, instead of at the extremes, results in further improvement.

In normal circuits, if realizable values for $\alpha_{y'}$ and $\alpha_{x'}$ are not obtained, it will be found that some of the individual coefficients are unnecessarily high. Best results usually accrue when the "uncorrected" drift has been reduced to a minimum by careful selection and design, so that individual components have low coefficients, although there are often limits to such a procedure.

For instance, in the adjustment of the temperature coefficient of an inductance, zero or a very low value is obtained only by cancellation of certain components of the coefficient against others,[5] and this practice may be no better, and is often worse than correcting for the original inductance coefficient by adjustment of the capacitance coefficient. A cyclic and reproducible coefficient is to be preferred to one which, though lower, is erratic.

Equation (18) is obtained in similar fashion to (17) and gives

$$\alpha_{z'} = \frac{2C_H}{C_z(C_{\max}^2 - C_{\min}^2)}\left(\frac{C_{\max}^2\cdot\theta_H}{n} - \frac{k\cdot C_{\min}^2\cdot\theta_L}{m}\right) + \alpha_z \tag{18}$$

and

$$\alpha_{y'} = \frac{2C_H\cdot C_y}{C_{\max}^2 - C_{\min}^2}\left(\frac{k\cdot\theta_L}{m} - \frac{\theta_H}{n}\right) + \alpha_y. \tag{18a}$$

Practical examples of the use of these expressions are shown in Fig. 5, the lower curves giving the drift prior to correction, and the upper ones the most probable value of the corrected drift. The lower curves have been assumed linear, as observations were made only at the frequency extremes.

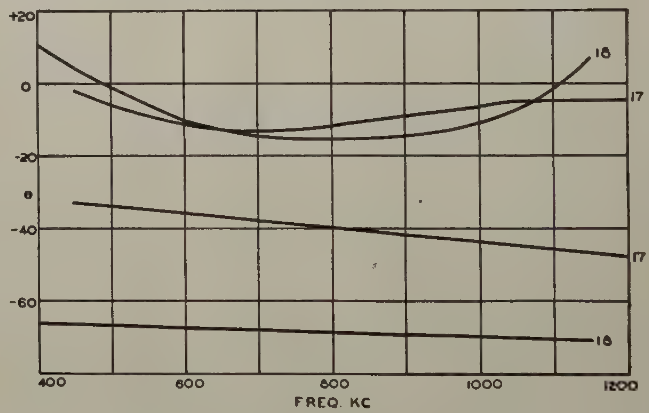

Fig. 5—Illustrating practical examples of drift correction by use of the padded circuit with coefficient tracking.

The upper curve for (17) gives a very good experimental check, since the exact values of coefficients required by the expressions were not readily obtainable. Values somewhat lower were used, and it was expected that the drift would be a few parts per million negative throughout the range. The inability to obtain the exact values for the coefficients, as determined by the expressions, will sometimes be the limiting factor in design.

The circuit used for (18) covers a frequency range of 2.9 to 1. In this case θ was made zero at two frequencies within the range, so as to reduce the drift at 800 kilocycles.

The examples are representative of what may be done with the use of stock broadcast receiver parts, having existing tolerances on the various electrical

[8] A. G. Manke, *RMA Eng.*, vol. 2, p. 18; 1938.

values and coefficients. It should be possible to have the average value of the corrected drift not more than ten parts per million per degree centigrade throughout a range of nearly 3 to 1.

Further improvement results from the restriction of the frequency coverage, and/or the use of components in which thermal drift has been a specific factor in design.

The adjustment of C_x and C_z was tried in a few cases, and it was found that the correction of such circuits called for coefficient values in excess of those possible. In any case, such expressions would have only limited application, since usually only one or the other is present in a circuit.

IX. Three-Point Tracking

The possibilities of three-point tracking have been briefly examined. It is evident that by operating on all three coefficients (α_x, α_y, and α_z), a third expression for θ at some other frequency could be set up and equated to zero. This would give a third tracking point at any suitable frequency within the range.

Such a third tracking point would be useful if there is considerable curvature in the two-point characteristic, as, e.g., in the examples of Fig. 5. Three-point tracking also gave unattainable values for $\alpha_{x'}$ and $\alpha_{z'}$, and it is not, in general, practically possible to have $\theta=0$ at more than two frequencies in the range, so that (17) or (18) gives the best possible compensation for any circuit.

The alteration of the third coefficient will have some effect on the shape of the curve, and may therefore give a smaller departure from zero at mid-frequencies.

IX. Conclusions

In circuits intended for operation at one frequency, thermal drift can be eliminated by adjustment of the temperature coefficient of a portion of the capacitance, as determined by (7).

In variable-capacitance-tuned circuits, the integrated drift is a minimum when $\theta_{L'}^2+\theta_{H'}^2$ is a minimum, the design information being given by (10). A close approximation to optimum conditions obtains when $\theta_{H'}$ is made zero, either by the use of (10c) or experimentally.

Circuits having θ less than ten parts per million per degree centigrade over normally used frequency ranges, are sometimes possible by the complementary adjustment of the coefficients of inductance and capacitance, as given in (11) and in the graphical construction of Fig. 3.

Variable-inductance-tuned circuits are a special case of the general formulas obtained for the variable-capacitance-tuned circuits. Minimum integrated drift occurs when $\theta_{L'}+\theta_{H'}=0$. These circuits have the advantage of being very suitable for "warm-up" drift reduction.

The padded circuit of the superheterodyne local oscillator is peculiarly adapted to thermal-drift correction by reason of its complex capacitance network. Suitable adjustment of two of the condenser coefficients, giving two-point coefficient tracking, results in a very small drift throughout the range, when the conditions of (17) or (18) are met.

Experimental work has confirmed the correctness of the expressions in so far as the experimental errors permit. It appears that the average value of the drift factor in the padded circuits is readily reduced to ten parts per million per degree centigrade for frequency ranges of up to 3 to 1.

Three-point tracking of the temperature coefficient of frequency, by the adjustment of three condenser coefficients, has been found to be impractical, due to the large values of coefficient called for in normal circuits.

Acknowledgment

Thanks are due to Amalgamated Wireless (Australasia), Ltd., for permission to publish this paper, to D. H. Connolly for useful comment in the draft stage, and to P. V. Moran for much of the checking and experimental work.

Attenuation of Electromagnetic Fields in Pipes Smaller Than the Critical Size*

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Summary—*A theoretical and experimental discussion is given of electromagnetic fields in pipes smaller than the critical size, especially with regard to attenuation, and based upon wave-guide theory. It is shown that the rate of attenuation, as the wavelength increases and passes through the critical value, approaches a high asymptotic value. Confirmatory experimental data are given. Simple formulas for attenuation are included.*

RECENTLY several papers[1–3] have appeared in which the propagation of electromagnetic waves in pipes has been discussed. These have indicated that for a given frequency there is a critical pipe size below which wave propagation is not possible, or in other words, for a pipe of a given size there is a cutoff frequency f_0 below which wave propagation does not occur. Aside from this, so far as the writer is aware, no discussion has previously been made of phenomena below cutoff or very near to it. In fact most of the present published theory is not valid for frequencies very near to f_0. It is the purpose of the present paper to discuss this case somewhat more completely.

The general expressions for the fields in cylindrical tubes may be written in the form[1,2]

$$\left.\begin{aligned} E &= E_1(r, \theta) \exp(jwt \pm \gamma z) \\ H &= H_1(r, \theta) \exp(jwt \pm \gamma z) \end{aligned}\right\}, \tag{1}$$

whence it is seen that the field along the axis of the tube is controlled by the factor $\exp(jwt \pm \gamma z)$, where γ, the propagation constant, may be written $\gamma = \alpha + j\beta$. The real part of γ is the attenuation constant and the imaginary part is the phase constant.

The forms of propagation constant discussed in the above references[1–3] are not valid for $f \leqq f_0$, except in Barrow's paper. He gives a form of γ which is valid for the E-type wave, over the whole frequency range, but does not discuss it, putting it immediately into a simpler form which does not hold over the whole range. Barrow's general expression is limited only by the assumption that the conductivity of the tube is very large but finite. After making several changes of notation it may be written

$$\gamma = \frac{2\pi}{\lambda_0}\left\{\left[1 - \left(\frac{\lambda_0}{\lambda}\right)^2 - \frac{\lambda_0^2 w}{4\pi^2}\right] + j\,\frac{\lambda_0^2 w}{4\pi^2}\right\}^{1/2}, \tag{2}$$

where λ_0 is the free-space wavelength corresponding to f_0, i.e., the cutoff wavelength, and

$$w = \frac{\sqrt{2}\,\omega\epsilon_1}{a}\sqrt{\frac{\omega\mu_2}{\sigma_2}}, \tag{3}$$

where

ϵ_1 = dielectric constant of the medium ($10^{-11}/36\pi$ farad per centimeter for air)

μ_2 = permeability of the tube (henries per centimeter)

σ_2 = conductivity of the tube (mhos per centimeter)

a = radius of the tube.

For copper pipes of practical sizes, the w term is negligibly small except very near cutoff.

The real part of (2) determines the attenuation. It is of interest to consider its variation as λ varies from below to above λ_0. This has been plotted in Fig. 1,

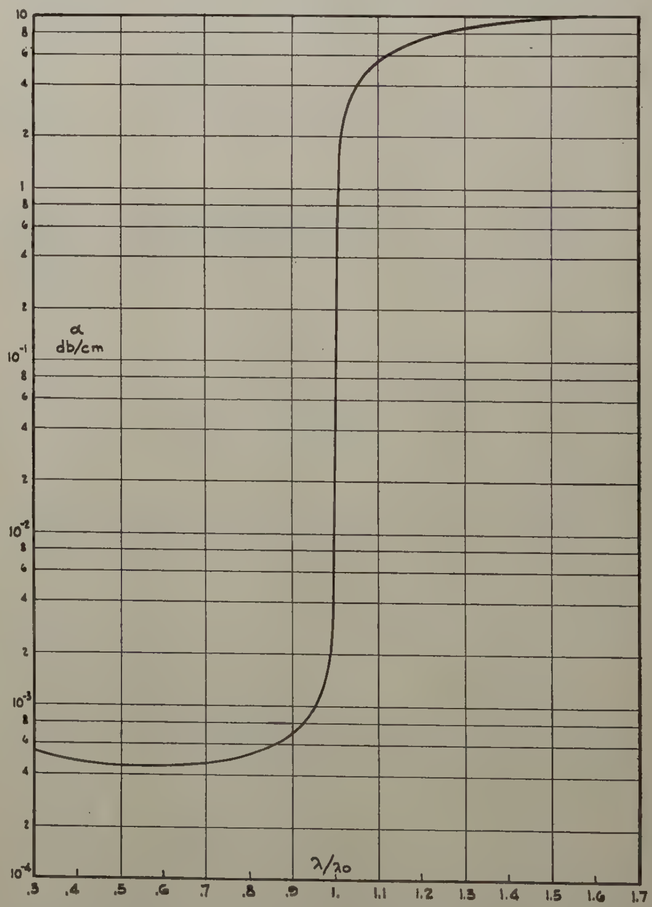

Fig. 1—Attenuation for the case of an $E_{0,1}$ type of field in a copper pipe of 1.58 centimeters radius, showing the variation through the cutoff point.

for the case of $E_{0,1}$ type of field in a copper tube of radius $a = 1.58$ centimeters. For $\lambda < \lambda_0$ the curve is identical with those published previously[1–3], except as

* Decimal classification: R110. Original manuscript received by the Institute, April 6, 1942.

† RCA Laboratories, Princeton, New Jersey.

[1] W. L. Barrow, "Transmission of electromagnetic waves in hollow tubes of metal," Proc. I.R.E., vol. 24, pp. 1298–1329; October, 1936.

[2] J. R. Carson, S. P. Mead, and S. A. Schelkunoff, "Hyper-frequency wave guides," *Bell Sys. Tech. Jour.*, vol. 15, pp. 310–333; April, 1936.

[3] G. C. Southworth, "Hyper-frequency wave guides," *Bell Sys. Tech. Jour.*, vol. 15, pp. 284–309; April, 1936.

$\lambda \to \lambda_0$ the attenuation does not become infinite as previously indicated, but rises very steeply to a comparatively high value, and for $\lambda > \lambda_0$ it approaches an asymptotic value of 13.2 decibels per centimeter.

The imaginary part of γ determines the form of the field along the z axis. For $\lambda < \lambda_0$, it is seen from (2), the imaginary part is very large compared to the real part; i.e., $\beta \gg \alpha$. Hence we see from (1),

$$E = E_1(r, \theta)\exp(-\alpha z)\exp(-j\beta z),$$

that the field is simple harmonic with small attenuation. However, for $\lambda > \lambda_0$, we have $\beta \ll \alpha$. The factor $\exp(-j\beta z)$ is very near unity. The harmonic variation disappears and the field decreases at a high exponential rate as determined by $\exp(-\alpha z.)$

At the cutoff point $\lambda = \lambda_0$ and $\gamma = \sqrt{w(-1+j)} = \sqrt{w}(\cos 67.50 + j \sin 67.50)$. Thus α and β are of similar magnitude.

It is evident that there is a continuous transition through the cutoff point, from a slightly attenuated simple harmonic wave to a highly attenuated exponential field.

It is possible to derive very simple expressions for the attenuation which are accurate except close to λ_0. To do this write γ^2 in the form

$$\gamma^2 = \rho\epsilon^{2\phi},$$

whence

$$\gamma = \sqrt{\rho}\,\epsilon^{\phi},$$

$$\alpha = \sqrt{\rho}\cos\phi,$$

and

$$\beta = \sqrt{\rho}\sin\phi,$$

where, from (2)

$$\rho = \frac{4\pi^2}{\lambda_0^2}\sqrt{\left(1 - \frac{\lambda_0^2}{\lambda^2} - \frac{\lambda_0^2 w}{4\pi^2}\right)^2 + \left(\frac{\lambda_0^2 w}{4\pi^2}\right)^2}.$$

Except very near to λ_0 the w term is negligibly small and this is closely approximated by

$$\rho = \frac{4\pi^2}{\lambda_0^2}\left(\frac{\lambda_0^2}{\lambda^2} - 1\right), \quad \text{for} \quad \lambda < \lambda_0,$$

and

$$\rho = \frac{4\pi^2}{\lambda_0^2}\left(1 - \frac{\lambda_0^2}{\lambda^2}\right), \quad \text{for} \quad \lambda > \lambda_0.$$

Consider the two cases:

1. $\lambda < \lambda_0$.

Here γ^2 has a large negative real part, and a small positive imaginary part, hence 2ϕ is almost 180 degrees and ϕ is almost 90 degrees. Hence we may write $\cos\phi = w/2\rho$, and $\sin\phi = 1$.

Therefore,

$$\alpha = \sqrt{\rho}\cos\phi = \frac{\lambda_0 w/4\pi}{\sqrt{\dfrac{\lambda_0^2}{\lambda^2} - 1}},$$

or, multiplying numerator and denominator by λ/λ_0,

$$\alpha = \frac{\lambda w/4\pi}{\sqrt{1 - \dfrac{\lambda^2}{\lambda_0^2}}}, \tag{4}$$

and

$$\beta = \frac{2\pi}{\lambda}\sqrt{1 - \frac{\lambda^2}{\lambda_0^2}}. \tag{5}$$

These are identical with the expressions derived by Barrow;[1] and Carson, Mead, and Schelkunoff[2], for this case.

2. $\lambda > \lambda_0$.

Here γ^2 has a large positive real part, and a small positive imaginary part. Hence, 2ϕ is nearly zero, $\cos\phi = 1$, and $\sin\phi = w/2\rho$. Therefore,

$$\alpha = \frac{2\pi}{\lambda_0}\sqrt{1 - \frac{\lambda_0^2}{\lambda^2}}, \tag{6}$$

and

$$\beta = \frac{\lambda_0 w/4\pi}{\sqrt{1 - \dfrac{\lambda_0^2}{\lambda^2}}}. \tag{7}$$

It is of interest that the attenuation in this case depends only on λ and λ_0 and is independent of the tube material, except for the assumption that it is of high conductitivy.

The above discussion applies only to E waves, since the basic expression (2) was derived for that case only. The case of H waves for $\lambda < \lambda_0$ has been discussed in detail by Barrow; Carson, Mead, and Schelkunoff, etc. The case $\lambda > \lambda_0$ may be handled by noting that the propagation constant for H waves may be written[4]

$$\gamma = \frac{2\pi}{\lambda_0}\sqrt{1 - \frac{\lambda_0^2}{\lambda^2}},$$

except for λ very near to λ_0. For $\lambda > \lambda_0$ this is real and thus represents the attenuation, which is seen to be identical with (6). The assumptions are the same in both cases. Hence (6) is valid for both E waves and H waves.

Equations (4) and (5) have been discussed in previous publications,[1-3] and will not be dealt with further here. Equations (6) and (7) however appear not to have been mentioned previously.[5] Equation (6) has been checked over the wavelength range from 7 to 10 centimeters for a tube of 1.58 centimeters radius, and for a field configuration of the $H_{1,1}$ type. These data, which were obtained by G. Fernsler, are plotted

[4] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Company, New York, N. Y., 1941, p. 539.

[5] R. A. Braden of this laboratory first noticed that our experimental measurements of attenuation rate followed an empirical law similar to equation (6). The similarity of this to expressions occurring in wave-guide theory led to the present investigation.

in Fig. 2. There is agreement within the accuracy of the measurements.

The extension or fringing of fields into tubes has been discussed in a number of instances[6,7] for static or low-frequency fields. These represent limiting cases given by $\lambda \rightarrow \infty$, and for which, from (6),

$$\alpha = 2\pi/\lambda_0. \tag{8}$$

Harnett and Case[7] have described three examples, to which this result is applicable, in which tubes are used

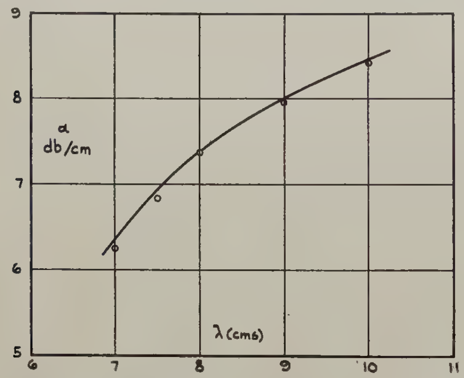

Fig. 2—Attenuation for the case of an $H_{1,1}$ type of field in a pipe of 1.58 centimeters radius. The circles are experimental points. The line is computed from equation (6).

as attenuators for signal generators, and in which three different field configurations are employed. The first type has an input electrode consisting of a small circular disk centrally located in the tube. This produces a radial electric field similar to that of the $E_{0,1}$ wave. Hence from wave-guide theory, $\lambda_0 = 2.62a$. Therefore, from (8),

$$\alpha = \frac{2\pi}{2.62a} \text{ napiers per centimeter}$$

$$= \frac{20.9}{a} \text{ decibles per centimeter}$$

$$= 20.9 \text{ decibles per radius,}$$

which is in agreement with previous results.[6,7] The second type has an input electrode consisting of a coil whose axis is normal to the axis of the tube. This produces a field configuration similar to that of the $H_{1,1}$ wave. The limiting value of α is found to be 16.0 decibels per radius. The third type consists of an input electrode in the form of a coil whose axis coincides with that of the tube. The field is analogous to that of the $H_{0,1}$ wave, and the limiting attenuation is found to be 33.3 decibels per radius. These three cases are in agreement with the limiting values found by H. A. Wheeler, and given in the Harnett and Case paper.

The formulas for $\lambda > \lambda_0$ for these three cases may be written as follows:

$$E_{0,1} \text{ type, } \alpha = 20.9\sqrt{1 - \frac{\lambda_0^2}{\lambda^2}} \text{ decibels per radius,} \tag{9}$$

$$H_{1,1} \text{ type, } \alpha = 16.0\sqrt{1 - \frac{\lambda_0^2}{\lambda^2}} \text{ decibels per radius,} \tag{10}$$

$$H_{0,1} \text{ type, } \alpha = 33.3\sqrt{1 - \frac{\lambda_0^2}{\lambda^2}} \text{ decibels per radius.} \tag{11}$$

Formulas for other types of waves may be derived by inserting the proper value of λ_0 in (6).

The accuracy of (6) and (7), and also the derived formulas (9), (10), and (11) may be estimated as follows: From the accurate expression (2) we see that the terms involving w (which were neglected in obtaining (6) and (7)) affect the order of magnitude of the result only when

$$1 - \frac{\lambda_0^2}{\lambda^2} \doteq \frac{\lambda_0^2 w}{4\pi^2} \doteq 10^{-4}$$

for copper pipes. This may be written

$$\lambda^2 - \lambda_0^2 = 10^{-4}\lambda^2,$$

or

$$(\lambda - \lambda_0)(\lambda + \lambda_0) \doteq 10^{-4}\lambda^2,$$

or

$$(\lambda - \lambda_0) = \Delta\lambda \doteq \frac{10^{-4}\lambda^2}{2\lambda_0},$$

or

$$\Delta\lambda \doteq 10^{-4}\lambda_0.$$

Thus, λ must be within 0.01 per cent of λ_0 before the w term is equal in magnitude to the sum of the other terms.

[6] I. Langmuir and K. T. Compton, "Electrical discharges in gases, Part II," *Rev. Mod. Phys.*, vol. 3, pp. 212–213; April, 1931.

[7] D. E. Harnett and N. P. Case, "The design and testing of multirange receivers," Proc. I.R.E., vol. 23, pp. 578–594; June, 1935. The attenuation formulas given in this paper were derived by H. A. Wheeler, but his derivations have not been published.

Institute News and Radio Notes

WINTER CONFERENCES—1943

The Board of Directors of the Institute, co-operating in the program of reducing long-haul railroad passenger traffic, has acted to eliminate the usual three-day convention and exhibition of component radio parts in New York this forthcoming January. To replace it, the Board has named Thursday, January 28, 1943, as a nationwide Winter Conference Day, on which as many Sections of the Institute as can possibly do so will hold simultaneous technical meetings. In this ambitious plan to "bring the mountain to Mahomet," the Directors believe that the Institute will achieve many of the more important objectives of a national convention, and, if Section officers enter into the spirit of the idea, will arouse local interest to an extent not possible through gathering at one point.

In New York City, the Board has accepted the kind invitation of the American Institute of Electrical Engineers, which during the week of January 25–29 is holding its National Technical Meeting there, to join with that society on January 28. Morning and afternoon sessions of the I.R.E. will be open to its own members and also to registrants at the A.I.E.E. technical meeting. The A.I.E.E. has courteously invited I.R.E. members to its communication and industrial electronics sessions, particulars of which will appear in the January PROCEEDINGS. The day's activities will culminate in the joint A.I.E.E.-I.R.E. evening meeting, with an address on the subject of Ultra-High Frequencies by Dr. George C. Southworth of Bell Telephone Laboratories.

Technical Papers

It is expected that technical papers and material used at the New York Conference will be made available, upon request, for simultaneous presentation at other I.R.E. Sections on the date named. Steps are being taken to make possible the exchange of papers among Sections so that each may contribute and all may benefit. If desired by the authors, papers so contributed will be considered by the Papers Committee and Board of Editors as to subsequent publication in the PROCEEDINGS. Papers for presentation at the conference must be submitted to the Institute office, 330 West 42nd Street, New York, N. Y., not later than January 1, 1943.

Other Features

Other features on the New York Conference program are: The Annual Meeting of the Institute; induction of officers for the year 1943; presentation of the Medal of Honor; award of Fellowships and other special awards. There will be no social activities at the New York Conference and no ladies' program. There will be no meeting of the Sections Committee.

Board of Directors

The Board of Directors met on October 7, 1942. Those present were A. F. Van Dyck, president; C. C. Chambers, I. S. Coggeshall, W. L. Everitt, Alfred N. Goldsmith, editor; L. C. F. Horle, C. M. Jansky, Jr., F. B. Lewellyn, Haraden Pratt, treasurer; F. E. Terman, B. J. Thompson, H. M. Turner, H. A. Wheeler, L. P. Wheeler, and H. P. Westman, secretary.

Institute representatives at 76 colleges and universities were appointed to serve until June 30, 1943.

A new Bylaw to follow present Bylaw Section 45 was adopted and reads as follows:

"The Board shall make appointments to the following Committees: Annual Review, Electroacoustics, Electronics, Facsimile, Frequency Modulation, Radio Receivers, Radio Wave Propagation, Standards, Symbols, Television, and Transmitters and Antennas, each year between January 1 and May 1 and the terms of appointments shall be from May 1 of the year when the appointments are made until April 30 of the following year. Additional appointments may be made to fill vacancies or to care for special cases as the need arises, with the term of the appointment expiring April 30."

The report of a special committee appointed to present to the Board a report on the advisability of, and on ways and means of, changing the membership-grade structure to permit the Institute to serve better a larger and more diversified membership was considered. The Constitution and Laws Committee was directed to prepare the necessary changes in the Constitution and Bylaws to put the proposals of the Committee into effect.

The Executive Committee was directed to appoint a Conference Committee to arrange for a technical conference in New York City during the latter part of January, 1943. This conference, which will be of one day's duration, will replace the normal Winter Convention which was canceled some months ago.

The regular November meeting of the Board of Directors was held on the fourth of the month and was attended by A. F. Van Dyck, president; Austin Bailey, C. C. Chambers, I. S. Coggeshall, W. L. Everitt, Alfred N. Goldsmith, editor; C. M. Jansky, Jr., F. B. Llewellyn, Haraden Pratt, treasurer; F. E. Terman, B. J. Thompson, H. M. Turner, L. P. Wheeler, and H. P. Westman, secretary.

The report of the Tellers Committee was received and the following new officers were declared elected:

President, 1943
L. P. Wheeler
Vice President, 1943
F. S. Barton
Directors, 1943–1945
W. L. Barrow
F. B. Llewellyn
H. A. Wheeler

On recommendation of the Awards Committee the following members will be advanced to Fellow at the time of the Annual Meeting which will be held in New York City in January: Andrew Alford, I. S. Coggeshall, J. B. Dow, L. E. DuBridge, P. C. Goldmark, D. E. Harnett, D. D. Israel, A. G. Jensen, G. F. Metcalf, and Irving Wolff.

The membership of the Appointments Committee was decided on and it will be comprised of L. P. Wheeler, chairman;

W. L. Everitt, F. B. Llewellyn, B. J. Thompson, and H. A. Wheeler.

A new Bylaw was adopted to follow the present Bylaw Section 44 and reads as follows:

"The Board of Directors is authorized to waive, in whole or in part, the application in any particular case of the contents of Bylaw Section 44 during the period ending December 31, 1943."

Section 44 of the Institute Bylaws prescribed the minimum requirements for the operation of sections. This Bylaw was adopted in anticipation that some sections may have difficulty in maintaining these minimum requirements basically as a result of general conditions and not through lack of interest or initiative on the part of the management groups.

A New York Section having been established, no further need exists for the New York Program Committee which formerly prepared the programs for New York meetings. Accordingly, Section 45 of the Institute Bylaws was amended to delete from the list of standing committees the name of the New York Program Committee.

Alfred N. Goldsmith was named to serve as a representative of the Institute on the Standards Council of the American Standards Association for the period 1943–1945, the Secretary being designated to serve as alternate for the same period.

Professor Everitt announced that there were no plans being made for a Broadcast Engineering Conference at Ohio State University during 1943.

The report of the Constitution and Laws Committee on proposed revisions of the Constitution and Bylaws submitted in accordance with the instructions issued at the previous meeting of the Board of Directors, was considered. A number of modifications were made in the proposals and the report returned to the Committee for further consideration by it.

Executive Committee

A meeting of the Executive Committee was held on September 18 and was attended by A. F. Van Dyck, chairman; Alfred N. Goldsmith, editor; Haraden Pratt, treasurer; B. J. Thompson, and H. P. Westman, secretary.

A memorandum on broadening the scope of the Institute was prepared for submission to the Board of Directors.

W. B. Cowilich was employed to serve as Assistant Secretary.

On October 2, A. F. Van Dyck, chairman; I. S. Coggeshall, Alfred N. Goldsmith, editor; R. A. Heising (guest); F. B. Llewellyn, Haraden Pratt, treasurer; B. J. Thompson, and H. P. Westman, secretary, attended a meeting of the Executive Committee.

It was agreed that the memorandum on "Broadening the Scope of the Institute" which was prepared at the previous meeting of the Executive Committee be forwarded to the Constitution and Laws Committee together with certain other information and views for its information and guidance in preparing proposed modifications of the Constitution and Bylaws. The proposed changes were requested in time for consideration by the Board of Directors at its November meeting.

Approval was granted of 88 applications for Associate, 41 for Student, and 4 for Junior membership.

A cordial invitation from the Institution of Electrical Engineers of London for Institute members visiting England to make use of its library facilities and to attend its meetings was accepted with thanks. A reciprocal privilege is afforded to members of the I.E.E. on visit to this country to make use of such facilities of our Institute as may be useful to them.

On October 30, a meeting of the Executive Committee was held and was attended by A. F. Van Dyck, chairman; I. S. Coggeshall, Alfred N. Goldsmith, editor; R. A. Heising (guest); F. B. Llewellyn, Haraden Pratt, treasurer; B. J. Thompson, and H. P. Westman, secretary.

J. B. Atwood, Stewart Becker, J. F. Johnson, H. T. Maser, M. D. McFarlane, E. R. Piore, and C. H. Wesser were transferred to Member grade. F. B. Bramhall, D. D. Carpenter, A. V. Dubinin, D. H. Marathe, E. D. McArthur, M. S. Neiman, C. T. Scully, A. H. Simons, G. F. Van Dissel, C. M. Wallis, and Michael Wysotzky were admitted to Member.

Approval was granted of 112 applications for Associate, 99 for Student, and 3 for Junior grade.

The petition for the establishment of a New York Section which was received on October 7 and signed by 5 Fellows, 13 Members, and 64 Associates in good standing was approved.

A report of the Constitution and Laws Committee on proposals to modify the Constitution and Bylaws of the Institute was reviewed.

Books

A-C Calculation Charts, by R. Lorenzen

Published by John F. Rider Publisher, Inc., 404 Fourth Avenue, New York, N. Y. 160 pages. 144 charts. 9×12 inches. Price $7.50.

After a brief historical and explanatory introduction, this large volume comprises 144 full-page reactance charts like those commonly used by communication engineers. Each chart is one cycle square, printed in two colors. All charts are alike except for the numbering of the scales. They cover the range of 10 cycles to 1000 megacycles, 0.01 ohm to 10 megohms, 0.1 micromho to 100 mhos.

Each of the 144 charts is accompanied by two conversion scales as an aid in impedance computations. One is a reciprocal conversion between ohms and mhos. The other is a square or square-root conversion for solving right triangles. There are two additional charts, one giving the relation $Q=X/R$ or B/G and the other giving $Q=\tan\theta$.

This reference book is recommended for libraries and laboratories where many reactance computations in various frequency ranges have to be made with accuracy nearly as great as that of a ten-inch slide rule.

H. A. Wheeler
Hazeltine Service Corporation
Little Neck, L. I., N. Y.

Handbook of Technical Instruction for Wireless Telegraphists, by H. M. Dowsett and L. E. Q. Walker

Published by Iliffe and Sons, Ltd., Dorset House, Stamford Street, London, S.E. 1, England, Seventh Edition, 1942. 664 pages. 618 figures. $5\frac{1}{4}\times8$ inches. Price, 25 shillings.

This Handbook aims chiefly to provide instruction and reference material for seagoing operators and others, relative to the general practice of marine wireless communication, illustrated by apparatus developed by British wireless companies.

About one fourth of the Handbook is devoted to the basic theoretical considerations which are applicable in all fields of radio communications. This material is clearly presented and numerous mathematical formulas most likely to be useful to the operator in the field are included.

Circuit diagrams are presented for a great number of different types of British shipboard apparatus, including transmitters, receivers, direction finders, auto alarms, emergency sets, and lifeboat sets, making the Handbook a convenient reference source for anyone working in this field.

H. O. Peterson
R.C.A. Communications, Inc.
Riverhead, L. I., N. Y.

The Radio Amateur's Handbook (Nineteenth Edition), 1942

Published by the American Radio Relay League, Hartford, Conn. 446 pages +8-page index+96-page catalog section. 680 figures. $6\frac{1}{2}\times9\frac{1}{2}$ inches. Price $1.00.

This is the nineteenth edition of the Radio Amateur's Handbook. Its arrangement and selection of material reflects the experience gained in the production of eighteen preceding editions.

About one third of the Handbook is devoted to the presentation of the basic theory of the components used in radio communication. This section includes a good chapter on practical antenna design.

There is also a section relative to the practices of station operation and traffic handling.

About one third of the Handbook pertains to construction plans and data for transmitters, receivers, transceivers, modulation equipment and other items of general use in the amateur's station. A comprehensive selection of tube characteristics and miscellaneous data are also included.

The catalog section is of interest and value in that many of the component parts

illustrated may be applicable to the radio problems incidental to the war effort.

H. O. PETERSON
R.C.A. Communications, Inc.
Riverhead, L. I., N. Y.

Acoustic Design Charts, by Frank Massa

Published by the Blakiston Company, Philadelphia, Pa. 219 pages+xiv pages+8-page index. $6\frac{1}{2}\times9\frac{1}{2}$ inches. Price $4.00.

Those engaged in the design of electro-acoustical apparatus will welcome the publication of this volume. It comprises a collection of charts, mostly in the form of straight line graphs on log-log co-ordinates, of quantitative data pertaining principally to acoustics and to mechanical vibrating systems. In many cases families of curves are given, which permit the effect of varying the parameters of a system to be evaluated without computation. Even a mere inspection of the charts is instructive, in that the ranges of values to be encountered in practical problems are indicated. A convenient feature is the use of both English and metric units in appropriate cases; for example, the length of a tube as a function of its resonance frequency is given in feet, in inches, and in centimeters. In most cases the scales chosen permit a constant percentage accuracy of reading over the large range of values covered. While the charts are not large in size, the precision is sufficient for most preliminary design calculations.

A wide range of material is covered, most of which is particularly applicable to the design of loudspeakers and microphones. The charts are arranged in ten sections covering fundamental relations in plane and spherical sound waves; attenuation of sound and vibrations; acoustical elements; mechanical and acoustical vibrating systems, including clamped, stretched, and piston diaphragms; design factors involved in direct radiator and horn-type loudspeakers; sound reproduction in rooms and in free space; design of electromagnetic systems; and, miscellaneous other data of use to acoustical engineers. There is a complete index with adequate cross references.

In few branches of engineering other than acoustics is a knowledge of fundamental principles more essential to the safe application of data such as are presented here. It is, therefore, unfortunate that neither the equations of the graphs nor references locating them in the literature can be found anywhere in the book. This is an omission which it is hoped will be rectified in future editions of this very useful volume.

Academicians may argue that graphic aids such as these are crutches for mental cripples. To this reviewer, however, the Massa charts seem more like a comfortable wheel chair in which one may, with little effort, glide smoothly along paths formerly negotiated only with painful effort.

BENJAMIN OLNEY
Stromberg-Carlson Telephone
Manufacturing Co., Rochester, N. Y

Contributors

CLEDO BRUNETTI

Cledo Brunetti (A'37) was born on April 1, 1910, at Virginia, Minnesota. He received the B.E.E. degree from the University of Minnesota in 1932 and the Ph.D. degree in 1937. From 1932 to 1936 he was a Teaching Fellow in the department of electrical engineering at the University of Minnesota and from 1936 to 1937, an instructor. From 1937 to 1939 Dr. Brunetti was an instructor in electrical engineering at Lehigh University and from 1939 to 1941, assistant professor of electrical engineering. During the summers of 1939 and of 1940 he was a research associate at the radio laboratory, National Bureau of Standards, Washington, D. C. Since May, 1941, he has been a radio physicist at the National Bureau of Standards, Washington, D. C., working on problems in connection with the war effort. He is a member of Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

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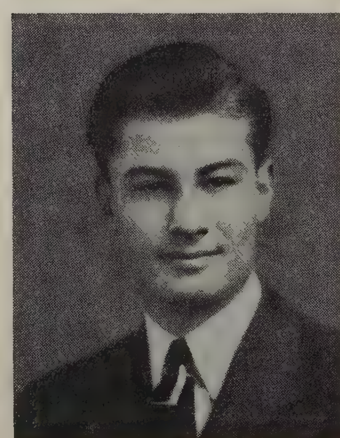
Beverly Dudley (J'24–A'27) was born on April 2, 1906, at Chicago, Illinois. He attended Armour Institute from 1926 to 1928. In 1935, he received the B.S. degree from Massachusetts Institute of Technology; from 1935 to 1939 he did graduate work in electrical engineering and physics at Columbia University. Mr. Dudley was assistant technical editor of *QST* at the American Radio Relay league from 1929 to 1930 and assistant secretary of the Institute of Radio Engineers from 1930 to 1932. He did technical editorial work at the General Radio Company during the summer of 1934 and experimental and factory production, test, and specifications on cathode-ray tubes and metal receiving tubes at the RCA Manufacturing Company, 1935–1936. Since 1936 he has been with the McGraw-Hill Publishing Company as assistant editor, associate editor, and managing editor of *Electronics*. From 1939 to 1941 he was managing editor of *Photo Technique*. At the present time he is the consulting editor for a new series of texts to be known as the Radio Communication Series. Mr. Dudley is the author of "Handbook of Photography; Making your Photographs Effective" and "Introduction to Radio Engineering," which is now in preparation. He is a member of the Optical Society of America and the Acoustical Society of America.

BEVERLY DUDLEY



C. HERBERT GLEASON

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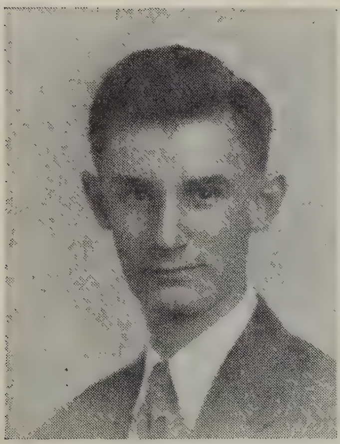
C. Herbert Gleason (A'42) was born at Halstead, Kansas on October 1, 1918. He received the B.S. degree in electrical engineering from the University of Kansas in 1940 and the M.S. degree in E. E. from the University of Missouri, 1942. He was a teaching assistant at the University of Missouri during 1940 and 1941 and an instructor at that University in 1941 and 1942. At present, Mr. Gleason is a junior



Leighton Greenough

engineer with the electronics engineering division of the Westinghouse Electric and Manufacturing Company, Bloomfield, New Jersey. He is a member of Sigma Xi and Tau Beta Pi.

Leighton Greenough was born on December 30, 1918. He received the B.S. degree in electrical engineering from Tufts College in 1940. In 1942, he received the M.S. degree in electrical engineering from Lehigh University, after two years as half-time Research Fellow. His work was with negative circuit elements. Since his graduation Mr. Greenough has been with the RCA Laboratories.

Ralph N. Harmon

Ralph N. Harmon (A '30) was born on August 29, 1906, at California, Pennsylvania. He received the B.S. degree in electrical engineering in 1928 from the Carnegie Institute of Technology. From 1925 to 1929, Mr. Harmon was with the Duquesne Light Company; since 1929, he has been with the Westinghouse Electric and Manufacturing Company.

E. G. Linder

Ernest G. Linder (A '40) was born in 1902 at Waltham, Massachusetts. He received the B.A. degree from the University of Iowa in 1925, and the M.S. degree in 1927. He was an instructor in physics at the State University of Iowa from 1925 to 1927; an instructor at California Institute of Technology from 1927 to 1928; a Detroit Edison Fellow at Cornell University from 1928 to 1932, where he received his Ph.D. degree in 1931. Dr. Linder was employed in the Research Division of the RCA Victor Company from 1932 to 1935; and the RCA Manufacturing Company, RCA Victor Division from 1935 to 1942. At present he is with the RCA Laboratories at Princeton, New Jersey. He is a member of the American Physical Society.

D. L. Waidelich (S'37–A'39) was born on May 3, 1915, at Allentown, Pennsylvania. He received the B.S. degree in electrical engineering in 1936 and the M.S. degree in 1938 from Lehigh University. From 1936 to 1938 he was a teaching assistant at Lehigh University; in 1938, he became an instructor at the University of Missouri and he is now an assistant professor of electrical engineering there. Mr. Waidelich is an associate member of the American Institute of Electrical Engineers and a member of Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

D. L. Waidelich

Standards on RADIO WAVE PROPAGATION

MEASURING METHODS

1942

Supplement to the PROCEEDINGS of the I. R. E.
Vol. 30, No. 7, Part II

THE INSTITUTE OF RADIO ENGINEERS

Standards

on

RADIO WAVE PROPAGATION

MEASURING METHODS

1942

The Institute of Radio Engineers, Inc.
330 West 42nd Street
New York, N. Y.

CONTENTS

INTRODUCTION

The Technical Committee on Radio Wave Propagation has formulated these standard measuring methods in frequent meetings and correspondence under the general guidance of the Standards Committee. Published with the approval of the Board of Directors, the report embodies the Institute's official recommendations to its members and the industry at large.

Suggestions and comments will be welcomed as an aid to committees preparing future reports. Correspondence should be addressed to the Institute of Radio Engineers.

CONCERNING THE INSTITUTE AND ITS STANDARDS ACTIVITIES

The Institute appointed its first standards committee in 1912, and the next year published a report dealing with definitions of terms, letter and graphical symbols, and methods of testing and rating equipment. Expanded reports appeared in 1915, 1922, 1926, 1928, 1931, and 1933, each of which combined, in a single document, data on all branches of the art.

Publication of the current series of standards, of which this one is a part, was begun in 1938.* It differs from earlier reports in that each individual booklet deals with a separate field. Under present policies, subdivision is being carried even farther and separate booklets are being issued in each field for definitions of terms, for symbols, and for measuring and testing methods.

Beginning with 1942, all standards are being published in the 8 1/2-×11-inch size to conform with the new format for the PROCEEDINGS of the I.R.E.

Distribution of Standards Reports

The Institute is mailing one copy of this report to every member who is in good standing for 1942.

Additional copies, if available, may be obtained by members or nonmembers from the Institute office at the price mentioned on the inside back cover.

Co-operation with Other Organizations

Throughout its life, the Institute has co-operated with other bodies in the establishment of standards. Last year, for instance, there were more than 50 official I.R.E. delegates to other standardization groups. The Institute is also the sponsor for the American Standards Association's Sectional Committee on Radio.

The Institute of Radio Engineers

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SECTION 1. METHODS OF MEASURING RADIO FIELD INTENSITY

1.1 GENERAL

Most measurements with which radio wave propagation is concerned involve the measurement of radio field intensity. Thus, the determination of power radiated from an antenna (Section 2) and the measurement of noise field intensity (Section 3) depend principally upon the measurement of field intensity. Standard methods for the measurement of this fundamental quantity are outlined here.

The general symbols and units used in this section are given in the following list. Where more specific or restrictive meanings are indicated by subscripts, etc., they are further defined after the equation in which used. In some cases a symbol definition is repeated where used, merely for convenience.

C = capacitance, in farads
d = distance from transmitter to receiver, in meters
E = radio field intensity, in volts per meter
f = frequency, in cycles per second
I = current, in amperes
l_L = effective length of loop antenna, in meters
l_D = effective length of dipole antenna, in meters
L_D = over-all physical length of a linear dipole antenna, in meters
l_V = effective length of grounded vertical antenna, in meters
L_V = over-all physical length of grounded vertical antenna, in meters
M = mutual inductance, in henrys
N = number of turns
P = power, in watts
Q = ratio of reactance to resistance in a resonant circuit
R = resistance, in ohms
S = area of loop antenna, in square meters
V = voltage, in volts
α = attenuator ratio (equal to or less than unity)
λ = wavelength, in meters

As may be seen in this list, all units used are those of the mks electromagnetic system. A person employing other units must reduce to these units before using the equations herein. For convenience in making such reductions a few relations are given here:

1 mile = 1609. meters
1 meter = 0.0006214 mile
1 square inch = 0.0006452 square meter
1 square meter = 1550. square inches

PREFIXES

micro = one millionth	deka = ten
milli = one thousandth	hecto = one hundred
centi = one hundredth	kilo = one thousand
deci = one tenth	mega = one million

Quantity	Name of mks unit	Number of cgs units in 1 mks unit
Length	meter	100
Area	square meter	10,000
Frequency	cycle per second	1
Voltage	volt	10^8
Radio field intensity	volt per meter	10^6
Current	ampere	1/10
Power	watt	10^7
Resistance	ohm	10^9
Conductivity	mho per meter	10^{-11}
Inductance	henry	10^9
Capacitance	farad	10^{-9}

1.11 THE TWO BASIC METHODS

Two general methods are applicable to the measurement of radio field intensity. One consists of measuring the voltage produced in a standard antenna by the field to be measured and computing the radio field intensity from the measured voltage and the dimensions and form of the standard antenna. The other consists of comparing voltages produced in an antenna by the field to be measured and by a standard field. The measuring sets for the two methods are of the same general form. For the standard-antenna method there are special requirements for the receiving antenna and the voltage-measuring equipment.

1.111 *Standard-Antenna Method*

The receiving antenna is of some standard form such that the voltage produced in it by a field of given intensity and polarization may be readily computed. The method of coupling the standard antenna to the voltage-measuring instrument is generally such that the voltage measured is not the voltage produced in the antenna by the field but bears a fixed ratio to it. This ratio (i.e., ratio of the voltage measured to the voltage produced in the antenna by the field) is called the voltage-transfer ratio; its magnitude is dependent upon the point of introduction of the calibrating voltage into the measuring circuit. Separate determination of it may be required.

A standard voltage source is provided, usually as part of the radio field-intensity-measuring apparatus, for calibrating the voltmeter. The term voltmeter is here used to mean the radio receiving portion of the equipment. The calibrating voltage may be fixed or it may be variable over a wide range. It may be inserted in series with the standard antenna or at some point in the coupling circuit, or it may be applied to the input

terminals of the voltmeter in place of the coupling circuit. When a fixed calibrating voltage is employed, the voltmeter is calibrated at a fixed sensitivity level, and a calibrated voltage attenuator is employed in conjunction with the voltmeter in order to measure voltages over the wide range required. Alternately, the voltage attenuator may be included in the standard voltage source; the calibrating voltage is then itself variable over a wide range and the voltmeter is calibrated at sensitivities corresponding to the voltages being measured. In either case, the measurement resolves itself into the comparison of two voltages, the known voltage and the voltage produced by the field, generally introduced at different points in the measuring circuit. The method used in determining the voltage-transfer ratio is generally dependent upon the particular form of antenna and the circuit arrangement employed.

The radio field intensity is computed on the basis of the antenna dimensions and form, the voltage-transfer ratio, the voltage from the standard voltage source, and the voltage-attenuator ratio.

1.112 *Standard-Field Method*

The standard field may be set up by a local transmitter. The field at the receiving antenna is computed from the dimensions of the transmitting antenna, its current and current distribution, the distance, and the effect of the ground. The measuring equipment associated with the receiving antenna generally consists of a sensitive vacuum-tube voltmeter employing double detection and with an indicator in the output of the second detector. While it is possible to adjust the standard field to equality with the field to be measured by varying the distance of the transmitter, it is usually more convenient to include an attenuator in the measuring equipment for comparing the voltages produced by the known and unknown fields and, hence, the two fields. When the angles of arrival of the standard and measured fields differ, a correction must be applied for any resultant difference in the voltages caused by directivity of the receiving antenna and effects of the ground.

1.12 SIMPLIFICATIONS AND PRECAUTIONS

The problem of radio field-intensity measurement is generally complicated by the varying nature of the field resulting from variability of the transmission medium, by the complex nature of the emission, and by the influence of the ground and of disturbing structures. Simplifications are introduced into measurement methods to keep the measuring apparatus from becoming unduly complex or unduly difficult to use or to facilitate analysis of the data.

In the case of ionospheric-wave fields, a really complete measurement would involve determination, from instant to instant, of the intensity of the vertical, longitudinal, and lateral electric- and magnetic-field components in each wave arriving at the receiving point; of the phase angles between the different components; and of the direction of arrival of each wave component. Fortunately, considerable practical information may be obtained by measuring only the vector components of the field which are used for communication and by treating the data statistically so as to determine over-all effects. Simple forms of receiving antennas thus become useful, the type of antenna used being dependent upon the component of the field. Statistical treatment of the data may be facilitated by employing a recorder in the radio field-intensity-measuring apparatus.

A further example of simplification is in the measurement of complex emissions. Thus, with frequency- or amplitude-modulated waves, it is desirable to remove the modulation so that a single-frequency field (the carrier) is measured. In emissions without carrier waves, such as a single-sideband wave, it is desirable to restrict the modulation to a single frequency, so that a single-frequency field is again measured. Occasions arise, however, when it is necessary to retain the normal modulation; for example, when considering the interfering effect of the over-modulation products of an amplitude-modulated wave or the interfering effect of a single-sideband emission.

In the case of complete modulation by telegraph-code signals, a good approximation may be obtained by using a voltage-measuring device with a perfectly linear and highly damped response, taking account of the fact that such a device will measure one half (=6 decibels below) the maximum or "mark" field, if the average keying is 50-per-cent marking.

In analyzing or reporting measured data, it is important to take into account the influence of the receiving site and surroundings (including topography), electrical properties of the ground, the proximity of disturbing structures, etc., and the orientation and height above ground of the receiving antenna. In view of the effect of ground reflection, the question arises as to whether the measured values of the resultant field should be reported or an attempt should be made to measure (or compute) the value of the incident field. In the case of a well-defined incident angle and when the electrical properties of the ground are accurately known, such a measurement would indicate the actual variation of the field at points equidistant from the transmitter without regard to the nature of the ground at various receiving points. In general, dealing with the incident field will eliminate confusion in the study of laws of attenuation and in the measurement of the directionl characteristics of antennas. However, the measurement of the incident field is always an indirect measurement involving a knowledge of the angle or angles of arrival and the electrical properties of the ground, and its proper determination thus introduces some difficulty.

A useful expedient under such circumstances is to use a horizontal doublet antenna for the measurement of resultant sky-wave fields, regardless of whether vertical or horizontal polarization is used. Practically, an equal amount of each type of polarization is present in the incident ionospheric-wave field for any condition of polarization at the transmitter. The advantage gained by receiving the resultant horizontal component is that the measured field is then practically independent of differences in the electrical properties of the ground at different receiving points. Hence, so long as the height of the receiving antenna above ground is stated or standardized, measurements (on a given frequency) at different sites may be directly compared.

In some cases, even though the resultant field is largely influenced by the nature of the ground, it should nevertheless be the field measured since it provides a more nearly correct picture of the actual conditions encountered in service. This is particularly true for measurements at standard broadcast frequencies (550 to 1600 kilocycles) where the antenna used by the average listener is principally responsive only to the resultant vertical component of the ionospheric waves.

1.2 STANDARD-ANTENNA METHOD

1.21 THE ANTENNA

The form of the standard receiving antenna which may be employed is governed by the polarization of the field component to be measured, by the nature of the field (i.e., high or low intensity, fading or steady, frequency, etc.), and by the degree of convenience of handling the antenna. The types generally employed are the loop antenna, the vertical- or horizontal-dipole antenna with a length of one-half wavelength or less, and the simple vertical antenna with output between one end and ground. Other types of antennas are frequently used, in which case they are calibrated in terms of one of the simpler antennas just enumerated or in terms of a standard field.[1,2]

The loop antenna may be used from the lowest frequency into the ultra-high-frequency range. Below about 3 megacycles, it is the practical equivalent of a vertical antenna (for low sky-wave angles), with the advantages of ease of computation of the voltage produced by a given field and convenience of handling (including the availability of a zero-response position). At higher frequencies, the effect of the response of the loop antenna to horizontal components of the electric field becomes more pronounced (since the loop antenna may then be at a substantial fraction of a wavelength above ground) so that the null becomes ill-defined, and there is increasing difficulty in defining the polarity of the component measured. Also, as the frequency is increased, there is increasing difficulty of securing a balanced connection of the loop antenna to the voltage-measuring set; moreover, because of the desirability under some circumstances for maintaining substantially uniform current distribution along the loop antenna, its pickup decreases rapidly with increasing frequency. At ultra-high frequencies, the loop antenna again becomes the practical equivalent of the vertical antenna, since low-angle reception is generally involved; however, technical difficulties in its use are materially increased.

[1] H. Pender and K. McIlwain, "Electrical Engineers Handbook," volume 5, "Electric Communication and Electronics," third edition, John Wiley, New York, N. Y., 1936.

[2] J. V. Cosman, "Portable field-intensity meter," *Communications*, vol. 18, pp. 22–23; September, 1938.

The shielded type of loop antenna has come into general use for direction finders and has been applied successfully in radio field intensity-measuring work. With this type of antenna, the loop-antenna circuit is generally not symmetrical with respect to ground, but it is desirable that the loop antenna shield be split at the top so that the shield is symmetrical with respect to ground.

The simple vertical antenna with output between one end and ground is conveniently applicable in the range from low frequencies up to about 30 megacycles. Care must be taken to make sure that the indicated voltage is not partly due to voltage induced in the ground system. A small symmetrical counterpoise ground is usually desirable.

The dipole type of antenna is generally applicable at frequencies above about 3 megacycles. It is particularly convenient at frequencies above 40 megacycles.

In addition to the foregoing, a type of antenna is occasionally used consisting of spaced parallel plates with a short conductor between centers.

Measurements of radio field intensity in established radio communication services are often made with more complex types of antennas or arrays. They afford the advantages of operating convenience and noise discrimination. The complex antenna is calibrated by comparison with either a loop antenna or a dipole antenna; however, this calibration may be of limited significance because of difference in the vertical directivity of the array and the comparison antenna.

1.22 CALIBRATION METHODS

In practice, a variety of circuit combinations are employed to carry out the standard-antenna method of radio field-intensity measurement. They are all essentially of the fundamental form previously described but differ primarily in the manner of calibrating the voltmeter.

1.221 *Loop-Antenna Calibration*

Figs. 1(a) and (b) illustrate two methods of loop-antenna calibration that have been used over the frequency range below 10 or 15 megacycles. In both, the

loop antenna is oriented and tuned for maximum output. The receiver attenuator is adjusted to give any convenient deflection on the output meter m and left at this setting. Then, with the loop antenna turned to a right-angle position so that the measured output is

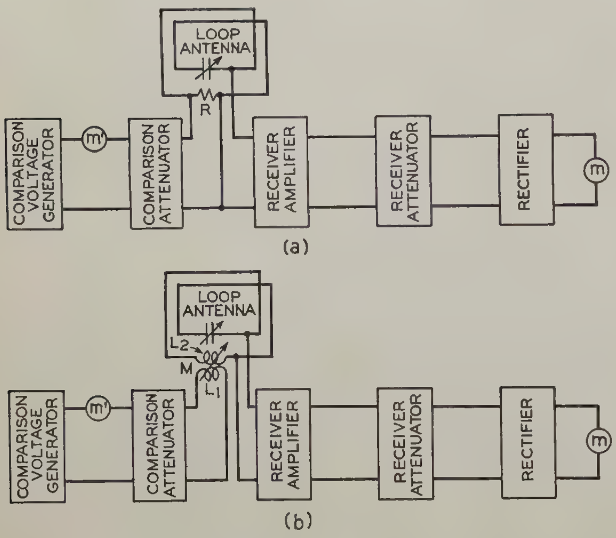


Fig. 1—Measuring sets using calibration input at level of field to be measured.

low, the comparison-voltage generator is turned on and tuned to approximately zero beat with the residual measured field, and the comparison-voltage attenuator is adjusted to give the same indication on meter m that was obtained when receiving the measured field. From the indication of meter m' and the comparison-attenuator setting, the current through the resistance R of Fig. 1(a) or inductance L_1 of Fig. 1(b) may be determined.

In the first case, when the natural resonance frequency of the loop antenna alone is at least two or three times that of the field measured, the measured radio field intensity E in volts per meter is

$$E = \frac{\alpha I R}{l_L} \tag{1}$$

where α = comparison-attenuator ratio (equal to or less than unity)

I = current at input to comparison attenuator, in amperes

R = resistance shown in Fig. 1(a), in ohms

l_L = effective length of loop antenna, in meters, defined as follows:

$$l_L = 2.094(10)^{-8} fSN \tag{2}$$

or

$$l_L = \frac{2\pi SN}{\lambda}. \tag{3}$$

For Fig. 1(b), with a similar precaution concerning design,

$$E = 2\pi \frac{\alpha f M I}{l_L} \tag{4}$$

where M = mutual inductance of coupled input, in henrys. Substituting from (2), (4) becomes

$$E = 3(10)^8 \frac{\alpha M I}{SN}. \tag{5}$$

In Fig. 1(a) and Fig. 1(b) the calibrating voltage injected into the loop antenna may in some instances be a fixed value. In such cases, the relationship between the field and calibrating voltages is determined by a calibrated attenuator in the receiver or by a calibrated output indicator.

Owing to difficulties that have been experienced with attenuator design at frequencies above a few megacycles, another method of calibration, illustrated schematically in Fig. 2(a), has been employed[3] extensively in the high-frequency range. The comparison-voltage attenuator is eliminated, and the receiver attenuator (which in the previous instances was not necessarily calibrated) is placed in the intermediate-frequency stages of a superheterodyne amplifier and accurately calibrated. At intermediate frequencies of the order of a few hundred kilocycles this is readily possible. To make a measurement, the loop antenna is oriented and tuned for maximum output, and the calibrated receiver attenuator is adjusted to give any convenient deflection on the meter m (Fig. 2(a)), then the attenuator ratio α is recorded, after which considerable attenuation is added and the comparison-voltage generator is turned on and tuned to the

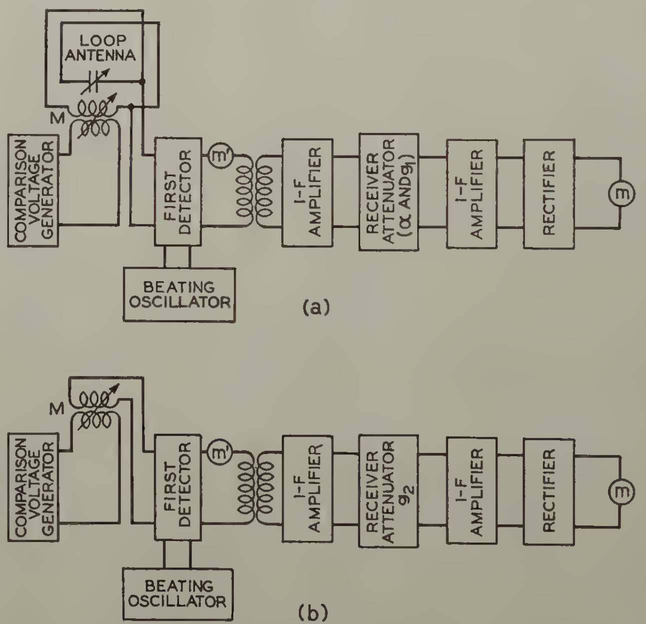


Fig. 2—Measuring set using high-level input for calibration.

frequency of the field to be measured. The beating oscillator is then turned off, and, using the first detector as a vacuum-tube voltmeter, the coupling M between the comparison-voltage generator and the loop antenna is adjusted to give V volts input as

[3] H. T. Friis and E. Bruce, "A radio field-strength measuring system for frequencies up to forty megacycles," Proc. I.R.E., vol. 14, pp. 507–519; August, 1926.

indicated by meter m' (Fig. 2(a)) in the first-detector plate circuit. The beating oscillator is then turned on, and the receiver attenuator is then adjusted to give the same indication of meter m that was used in the field measurement. The attenuator ratio is recorded and considered as "set gain" g_1.

Without changing the coupling M, a switch is thrown which effectively converts the first part of the circuit in Fig. 2(a) to that shown in Fig. 2(b). The voltage originally produced in the loop antenna is thus measured directly. The loop-antenna step-up or Q is normally so great that this measurement cannot be made on meter m'. It is determined by again adjusting the receiver attenuator to give the original field reading on meter m. This attenuator ratio may be called g_2. The radio field intensity in volts per meter is

$$E = \frac{g_1^2 V}{\alpha g_2 l_L} \tag{6}$$

where V = voltage input to first detector as read by meter m' (= voltage from comparison-voltage generator multiplied by half the loop-antenna Q factor)

α = receiver-attenuator ratio (equal to or less than unity) to give a convenient deflection of meter m for antenna voltage produced by the field

g_1 = receiver-attenuator ratio (equal to or less than unity) to give same deflection of meter m when V volts are applied to the input of the first detector

g_2 = receiver-attenuator ratio (equal to or less than unity) to give the same deflection of meter m when the voltage from the comparison-voltage generator is applied to the input of the first detector.

If desired, V may be made one volt, all attenuator ratios given in decibels, and the effective length of the loop antenna expressed as decibels above 1 meter as L_L, so that an expression for the field in decibels above 1 volt per meter in this case is

$$E_{\text{db}} = 2G_1 - A - G_2 - L_L \tag{7}$$

where all quantities are the equivalent in decibels of those represented by the small letters of (6); the letter A corresponds to α.

An important requirement in connection with the above method of measurement is that the amplification of the amplifier be the same for the high input used during calibration as it is when receiving weak fields. It is also important to avoid nonlinearity of the first detector.

In another method of calibration, the calibrating voltage is applied to the input of the vacuum-tube voltmeter and is adjusted to equality with the voltage developed across the input by the loop antenna. A measurement of the voltage step-up in the loop antenna is then required for computing the field voltage induced in the loop antenna. The radio field intensity is computed on the basis of the induced-field voltage and the antenna dimensions.

The method of measuring the voltage step-up factor is such as to avoid errors which arise from the distributed capacitance between turns of the loop antenna, when the calibrating voltage is inserted in the center of the antenna. The method utilizes a precision variable condenser of low capacitance connected in parallel with the tuning condenser. The condenser is set at its mid-value for loop-antenna resonance and is adjusted on either side of resonance until the voltage across the condenser drops to 0.707 times its value at resonance. Then, calling C_1 and C_2 the two corresponding capacitance values of the auxiliary condenser, and $\Delta C = C_2 = C_1$, the Q of the loop antenna may be shown to be very nearly equal to

$$Q = \frac{2C}{\Delta C}, \tag{8}$$

where C represents the total tuning capacitance at resonance.

Thus, if V is the voltage across the antenna-tuning condenser, the radio field intensity in volts per meter is

$$E = \frac{V}{l_L Q} = \frac{V \Delta C}{2 l_L C}. \tag{9}$$

It should be noted that V is the total voltage across the tuning condenser.

Generally speaking, the errors in the final result depend on four factors, namely, errors in computation of effective length, errors in determination of voltage-transfer ratio, errors in the calibrating current or voltage, and errors in the attenuating means. The error in the determination of voltage-transfer ratio is generally less than 3 per cent if the working frequency is greater than 2.5 times the natural frequency of the loop antenna.

The accuracy of measuring the calibrating current or voltage is dependent on the instrument used. If instruments of the rectifier type such as vacuum-tube voltmeters are used, errors of about 10 per cent (about 1 decibel) can result if the waveform of the voltage-generator output is not sinusoidal.

Errors in the attenuating means depend on design, construction, and calibration.

A loop antenna tuned to a frequency differing by a fixed amount from the operating frequency is particularly suitable for measuring high field intensities of the order of 10 millivolts per meter to 50 volts per meter. The Q factor of the loop antenna may be computed directly from the values of the operating frequency and the frequency to which the loop antenna is tuned. Furthermore, since the operation is considerably off resonance, the error in the determination of voltage-transfer ratio may be made negligible.

In this method the loop antenna is tuned to a known frequency f_1 which differs from the operating frequency f by at least 10 per cent. The calibrating voltage is applied directly to the input circuit of the voltmeter, replacing the loop antenna. A voltage attenuator operating at an intermediate frequency is used to compare the voltage delivered by the loop antenna (across one half the tuning condenser) with

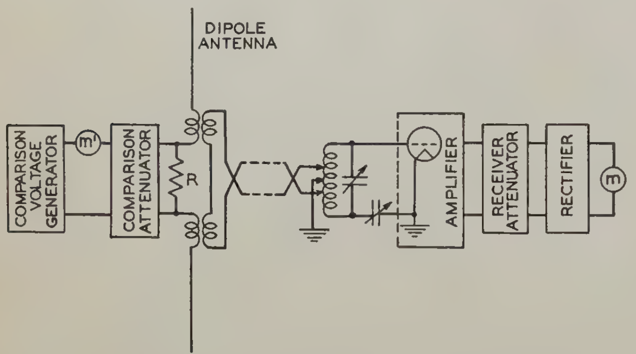


Fig. 3—Measuring set using dipole antenna.

the calibrating voltage. The voltage induced in the loop antenna by the measured field is computed from the measured voltage and a voltage step-up factor derived from the ratio of the operating frequency to the frequency to which the loop antenna is tuned, f/f_1. The radio field intensity is computed from the induced voltage and the dimensions of the loop antenna.

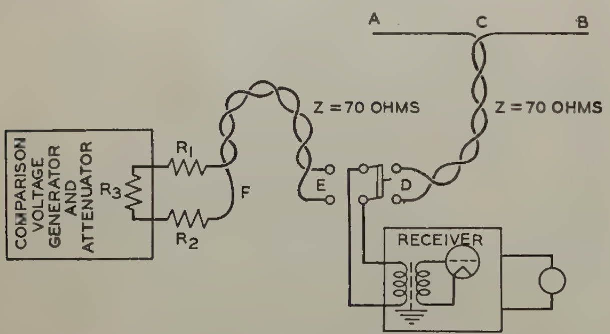


Fig. 4—Measuring set using dipole antenna.

The voltage step-up Q at the operation frequency f is

$$Q = \frac{f_1 Q_1}{2f\sqrt{1 + Q_1^2\left(\frac{f_1}{f} - \frac{f}{f_1}\right)^2}}, \tag{10}$$

where Q_1 is the voltage step-up at the frequency f_1 to which the loop antenna is tuned.

For $f/f_1 < 0.9$ and usual practical values of Q_1, the voltage step-up is given within 1 per cent by the simple expression

$$Q = \frac{1}{2(1 - (f/f_1)^2)}. \tag{11}$$

1.222 *Dipole-Antenna Calibration*

The methods of calibration used in a measuring set with a loop antenna may, in general, be applied to a half-wave dipole antenna.

One of the several possible direct methods of calibration is illustrated schematically by Fig. 3. The output of a comparison-voltage generator indicated on meter m' is attenuated by a known amount before passing through the resistance R. If the current through meter m' is I and the comparison-attenuator ratio α, the voltage in the dipole system is αIR. If, without disturbing the receiver, the comparison attenuator is adjusted to give the same deflection on the measuring-set output meter m as was obtained when receiving the field, then the radio field intensity in volts per meter is

$$E = \frac{\alpha IR}{l_D}, \tag{12}$$

where l_D is the effective length of the dipole antenna in meters. It is defined as follows:

$$l_D = \frac{L_D}{2} \frac{\tan \frac{\pi L_D}{2\lambda}}{\frac{\pi L_D}{2\lambda}}, \tag{13}$$

where L_D = the over-all physical length of the dipole antenna, in meters.

The effective length of a half-wave dipole antenna is, consequently, λ/π. When the dipole antenna is very short compared to the wavelength, the effective length is practically one half of the over-all physical length.

A calibration of this kind can be accomplished satisfactorily only when the measured field is absent, since the dipole antenna does not offer as convenient a means for suppressing the received field as a loop antenna that may be rotated. To use such a method in the presence of the measured field it is necessary to provide a known attenuation ratio of about 10 (i.e., 20 decibels) or more in the receiving circuit ahead of meter m and to arrange for this to be switched into the circuit when calibrating.

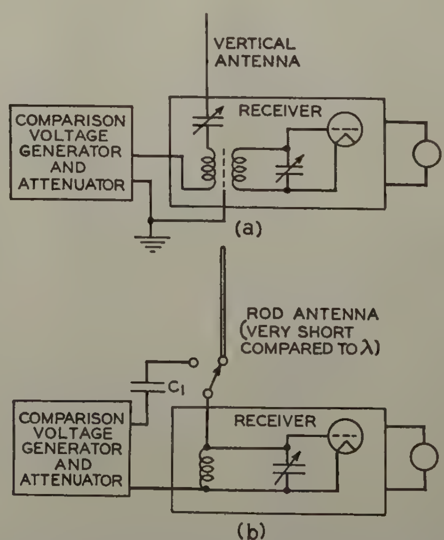
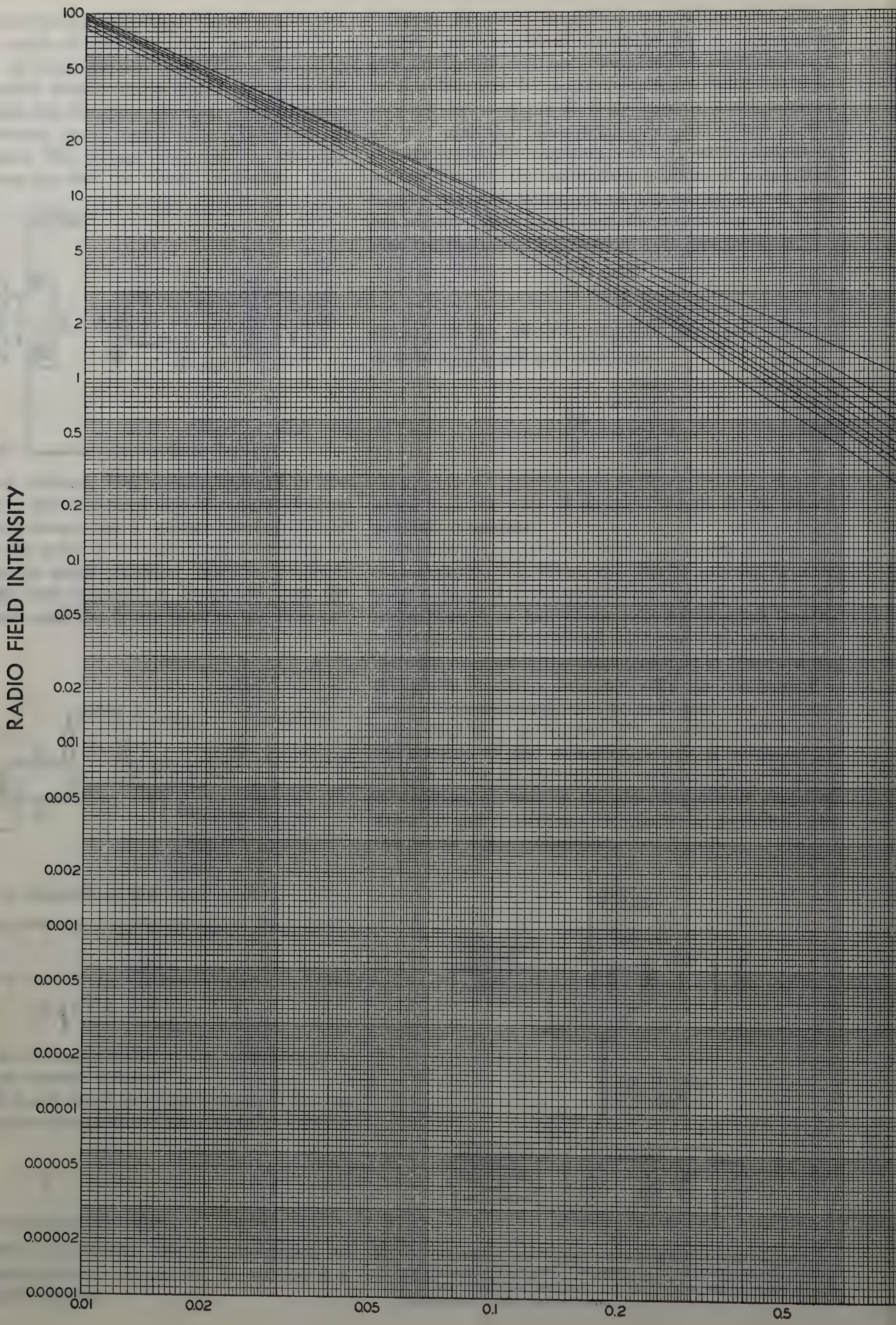


Fig. 5—Measuring sets using a vertical antenna with output from one end against ground. The setup is to be mounted on a tripod or other insulating support. In (b), C_1 is made equal to the capacitance of the rod antenna.

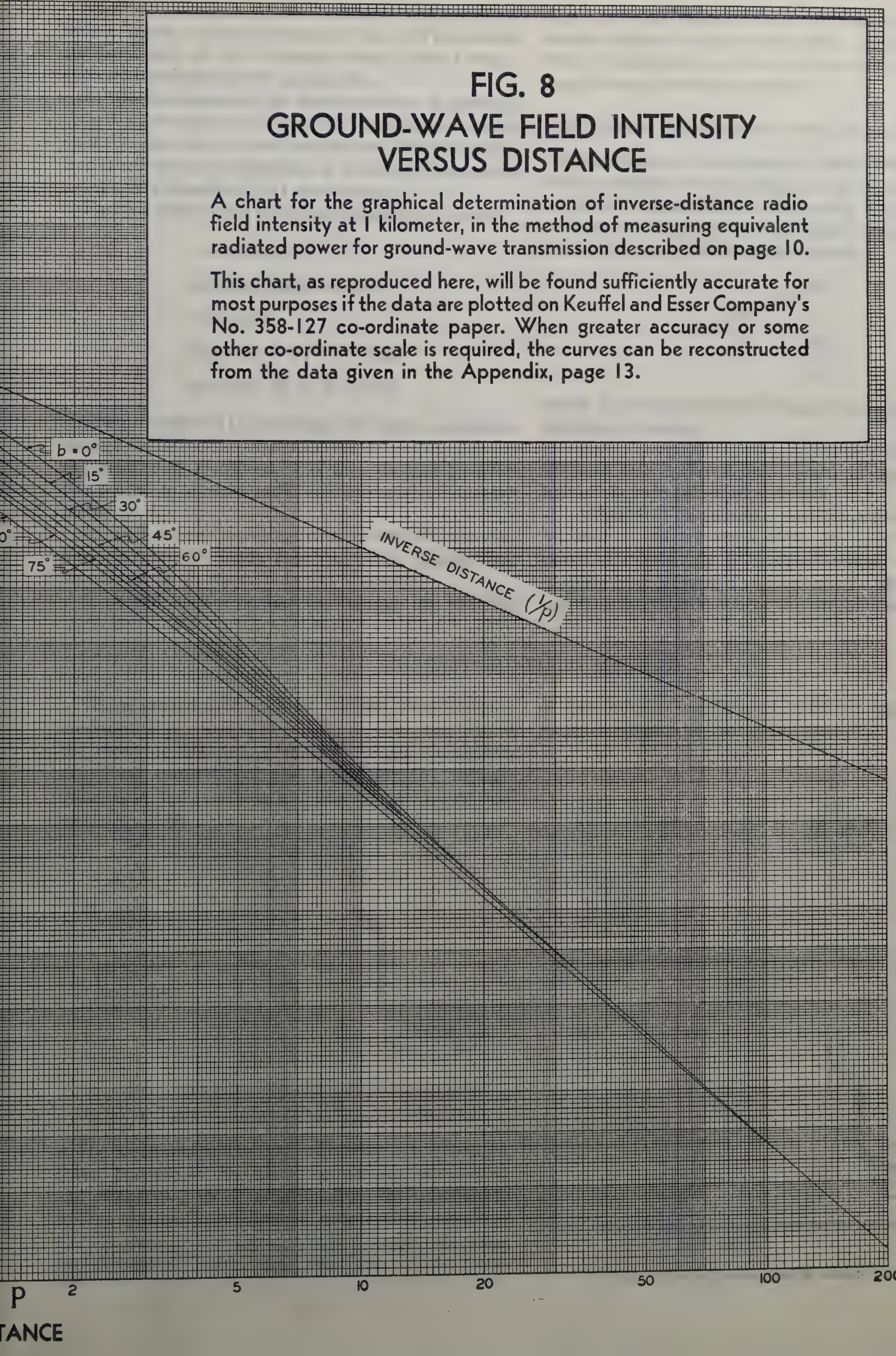


RADIO FIELD INTENSITY
100
50
20
10
5
2
1
0.5
0.2
0.1
0.05
0.02
0.01
0.005
0.002
0.001
0.0005
0.0002
0.0001
0.00005
0.00002
0.00001
0.01
0.02
0.05
0.1
0.2
0.5

FIG. 8
GROUND-WAVE FIELD INTENSITY VERSUS DISTANCE

A chart for the graphical determination of inverse-distance radio field intensity at 1 kilometer, in the method of measuring equivalent radiated power for ground-wave transmission described on page 10.

This chart, as reproduced here, will be found sufficiently accurate for most purposes if the data are plotted on Keuffel and Esser Company's No. 358-127 co-ordinate paper. When greater accuracy or some other co-ordinate scale is required, the curves can be reconstructed from the data given in the Appendix, page 13.



Another arrangement for making measurements with a dipole antenna is shown in Fig. 4. This system has been used in the frequency range from 4 megacycles to well above 100 megacycles.

In this arrangement the dipole antenna is operated at half-wave resonance. Generally it will have a length about 95 per cent of a half wavelength. For this condition it will have an impedance Z of about 70 ohms and an effective length of 0.31 wavelength.

For this system the following constants are used:

$$AB = 0.475\lambda$$

$$\text{effective length} = 0.31\lambda$$

$$EF = CD$$

$$R_1 + R_2 + R_3 = 70 \text{ ohms}$$

$$\text{generally } R_3 < R_1 \text{ and } R_2$$

The procedure is to determine the voltage-generator output V to give the same receiver output as given by the field being measured. V is then divided by effective length to give the radio field intensity in volts per meter.

In cases where the voltage drop in line CD is known, line FE may be made short and the voltage drop accounted for in the calculation of results.

It should be noted that the assumption that the dipole antenna impedance Z is 70 ohms is an approximation subject to appreciable error, if the dipole antenna is operated in a location less than a half wavelength from a reflecting surface.

1.223 *Calibration of Simple Vertical Antenna*

Arrangements for measuring radio field intensity with a simple vertical antenna with output between one end and ground are shown in Fig. 5 (a) and 5 (b).

The effective length of a grounded vertical antenna (in meters) may be defined as follows:

$$l_V = \frac{L_V}{2} \frac{\tan \dfrac{\pi L_V}{\lambda}}{\dfrac{\pi L_V}{\lambda}}, \tag{14}$$

where L_V = over-all physical length of grounded vertical antenna, in meters.

When the vertical antenna is very short compared to the wavelength, the effective length is practically one half of the over-all physical length.

In Fig. 5(b) is shown a method particularly applicable to rod antennas which are very short compared to wavelength. In this arrangement the receiver is calibrated through a capacitance C_1 which is made equal to the capacitance of the rod antenna.

1.3 STANDARD-FIELD METHOD

The standard-field method of radio field-intensity measurements consists of setting up a known field at the receiving antenna, generally through the agency of a local transmitter,[4] and comparing the voltages induced in the receiving antenna by the known and unknown fields.

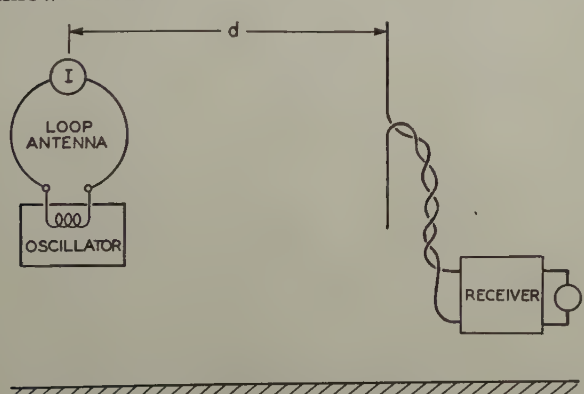


Fig. 6—Over-all calibration by field from a local transmitter.

The local transmitter generally employs a loop antenna, because of the ease in computing its radiation and of checking stray radiation from the transmitter. At frequencies above about 40 megacycles, some form of open antenna may be used, as for example the parallel-plate antenna or the rod antenna.

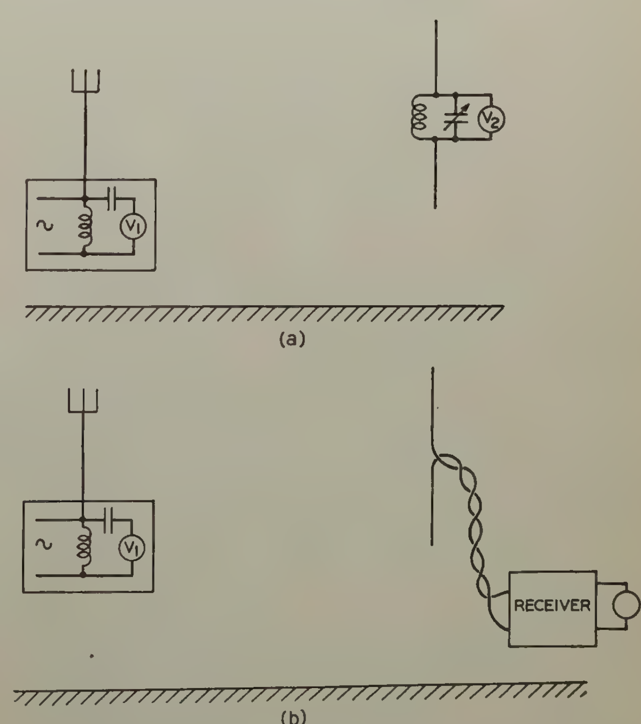


Fig. 7—Over-all calibration by radiated field.

[4] J. C. Schelleng, C. R. Burrows, and E. B. Ferrell, "Ultra-short-wave propagation," Proc. I.R.E., vol. 21, pp. 427–463; March, 1933.

The field at the receiving antenna is computed from the dimensions of the transmitting antenna, its current and current distribution, the distance, and the effects of the ground, taking into account the elevations of the transmitting and receiving antennas.

The method is illustrated in Fig. 6. At the higher frequencies it is generally possible to reduce the effect of ground to a small value by suitable choice of the elevations and distance apart of the transmitting and receiving antennas.

The radio field intensity at short distances from a radiating loop antenna in free space in volts per meter is

$$E = 1.316(10)^{-14}\frac{f^2NSI}{d}\sqrt{1+\frac{2.28}{f^2d^2}(10)^{15}} \qquad (15)$$

or

$$E = \frac{120\pi^2NSI}{\lambda^2 d}\cdot\sqrt{1+\left(\frac{\lambda}{2\pi d}\right)^2}. \qquad (16)$$

When the distance between the radiating loop antenna and the receiving antenna is a half wavelength, the radical expression becomes $\sqrt{1+0.101}$ which is about 1.05 (0.4 decibel above unity). Hence the second term increases the field to 1.05(=0.4 decibel above) the radiation field at this distance.

Another method of calibration using the field radiated by a local transmitter is shown in Fig. 7. In this system the local transmitter is provided with an instrument having considerable scale range and a reading proportional to antenna current. A vacuum-tube voltmeter V_1 may be used for this purpose. A range of several hundred to one (50 or 60 decibels) is readily obtainable in a multirange instrument. Transmitter output can be adjusted by variation of voltage or other means.

The first step is shown in Fig. 7 (a). Here the transmitter is set to radiate high power and the reading of V_1 observed. The field at a desired location is measured directly by means of a dipole and a simple voltmeter V_2. This operation gives radio field intensity for that value of V_1 for that location of receiving antenna.

The antenna of the system to be calibrated is then placed in that location (Fig. 7(b)) and the transmitter output reduced by an amount indicated by V_1. The radio field intensity is now at a value directly derived from the previous operation and may be used to calibrate the over-all sensitivity of the receiving system.

1.4 MISCELLANEOUS

1.41 ACCURACY OF MEASUREMENT

The precision attainable in practice depends not only upon the design of the equipment but upon the conditions under which the measurements are taken. The number of operating steps required in taking a reading, the number of instruments which must be read, and the number of adjustments to be made all have a bearing on the attainable accuracy. Assuming ideal conditions of location, weather, operation personnel, etc., an accuracy to within about 5 per cent (about ½ decibel) can be attained for frequencies below about 3 megacycles per second, and an accuracy to within about 10 per cent (about 1 decibel) for frequencies from about 3 to about 100 megacycles per second. The foregoing estimates are for a range of radio field intensities from about 100 microvolts per meter to about 1 volt per meter. Under usual operating conditions the accuracies may not be better than to within about 10 per cent (about 1 decibel) for frequencies below 3 megacycles per second and to within 40 per cent (3 decibels) for frequencies from 3 to 100 megacycles.

The attainable accuracies cited for ideal conditions appear to be adequate for practical purposes although improvement in the precision of measurements is always acceptable. It is desirable that the accuracy[5] of measurement of any data reported be stated.

1.42 AUTOMATIC RECORDING

In the measurement of fields of variable intensity, such as ionospheric-wave fields and disturbance fields, the accumulation of data is facilitated by the use of automatic recording methods. Arrangements for adapting the measuring apparatus to provide for recorder operation have been treated in a number of publications. The rectified output of the intermediate-frequency amplifier is used for automatic volume control of the amplifier and also, by direct-current amplification, for operating the recorder. Ordinary receiving equipment may also be adapted for radio field-intensity recording by connecting the recorder in the plate circuit of one or more of the tubes under automatic volume control. In either case, a substantially logarithmic relation between the recorder deflection and the radio field intensity may be obtained and is convenient in view of the great time variation of intensity often encountered.[6-10]

[5] Harry Diamond, Kenneth A. Norton, and Evan G. Lapham, "On the accuracy of radio field-intensity measurement at broadcast frequencies," *Nat. Bur. Stands. Jour. Res.*, vol. 21, pp. 795–818; December, 1938.

[6] W. W. Mutch, "A note on an automatic field-strength and static recorder," Proc. I.R.E., vol. 20, pp. 1914–1919; December, 1932.

[7] K. A. Norton and S. E. Reymer, "A continuous recorder of radio field intensities," *Nat. Broad. Sys. Jour. Res.*, vol. 11, pp. 373–378; September, 1933.

[8] R K. Potter and A. C. Peterson, Jr., "The reliability of

The time constant of the automatic-volume-control circuit will determine whether nearly instantaneous values[11] of the field or values representing short-time averages are obtained. The choice of the time constant will depend upon the ultimate use to be made of the measured data. It will be evident from the following remarks that a short time constant should be chosen, if the measured data are to be used in indicating the variations caused by the phase interference of waves arriving over various transmission paths; whereas a time constant long compared to the fading period may be chosen, if the measured data are to be used for showing the effects of variation in the transmission medium from hour to hour, day to day, season to season, etc.

A useful presentation of the measured data for fading fields is obtained by showing graphically the radio field intensities exceeded for various percentages of the time. It is obvious that for an accurate representation of the distribution of radio field intensities over a short time, such a graph must be determined from measurements of instantaneous radio field intensity. This graph will correspond to the Rayleigh distribution curve and will remain the same for time periods during which the layer absorption remains constant (i.e., up to about one hour). However, it can be shown that, if a graph representing the distribution of radio field intensities over a long time period is desired, accurate results will be obtained whether the graph is computed from instantaneous measurements or measurements which correspond to short-time averages of the instantaneous radio field intensities. This arises from the fact that the variations caused by changes in layer absorption are usually much larger and thus of greater influence than the short-time variations due to phase interference. An obvious advantage of using average values is that these may be obtained automatically by a proper choice of time constant in the measuring apparatus, and a material reduction in the work of analyzing the records is thereby obtained.

short-wave radio-telephone circuits," *Bell Sys. Tech. Jour.*, vol. 15, pp. 181–196; April, 1936.

[9] J. P. Taylor, "Graphic recording of field intensities," *Broadcast News*, vol. 23, pp. 6–9, 28, 33; December, 1936.

[10] Testimony of K. A. Norton, R. Bateman, and C. A. Ellert at Federal Communications Commission hearing on Ship Power, November 14, 1938. (F. C. C. mimeo 30539.)

[11] Lord Rayleigh, "On the resultant of a large number of vibrations of the same pitch and of arbitrary phase," *Phil. Mag.*, vol. 10, pp. 73–78; 1880. Also, "Theory of Sound," by Lord Rayleigh, second edition, paragraph 42a, Macmillan and Company, London, England, 1894. This problem is also treated in the "Collected Works" of Lord Rayleigh.

An integrating circuit of about 1 minute charge and 1 minute discharge time constants has been found useful for many purposes.

1.43 RECEIVING-ANTENNA POWER OUTPUT

When using certain types of antennas, a consideration of the relation between impressed radio field intensity and power output of the receiving antenna is sometimes of more practical use than a determination of open-circuit voltage. The receiver with a calibrated resonant circuit at its input then simulates a wattmeter rather than a voltmeter. The antenna is connected either directly or through a transmission line to this wattmeter, and, if the transmission loss is known, the power in the antenna may be determined. When the antenna impedance matches that of the connected circuits, half of the power is dissipated in the antenna and the remainder in the connected system including the wattmeter.

The maximum power output in watts of a receiving dipole antenna operated at half-wave resonance is, very nearly:

$$P = 3.12(10)^{13}\frac{E^2}{f^2} \tag{17}$$

or,

$$P = \frac{\lambda^2 E^2}{2885}. \tag{18}$$

1.44 PRESENTATION OF THE DATA OF MEASUREMENT

In view of the several factors which influence measured values of radio field intensity, information pertaining to the following should be provided to render measurement data generally useful:

(a) statement of which vector component of the field is measured;

(b) complete description of the measuring location and antenna setup, including topography, electrical properties of the ground, the proximity of disturbing structures, orientation of the receiving antenna and its height above ground, etc.;

(c) detailed description of the measuring equipment and procedure;

(d) in the case of ionospheric-wave fields, the time distribution of instantaneous or short-time-averaged radio field intensities throughout the time occupied by the measurements;

(e) for measurements made in the proximity of the ground, statement of which field is measured, i.e., incident or resultant;

(f) estimated accuracy of the measurements.

1.5 ADDITIONAL REFERENCES

A. A. Pistolkors, "The radiation resistance of beam antennas," Proc. I.R.E., vol. 17, pp. 562–579; March, 1929.

P. S. Carter, "Circuit relations in radiating systems and applications to antenna problems," Proc. I.R.E., vol. 20, pp. 1004–1041; June, 1932.

S. Goldman, "Dipoles and reflectors, a short review," *Electronics*, vol. 13, pp. 20–22; May, 1940.

SECTION 2. METHODS OF MEASURING POWER RADIATED FROM AN ANTENNA

2.1 GENERAL

This section outlines briefly methods of measuring the radiated power of an antenna and describes the important considerations for securing useful and accurate measurements. In all the methods, radiated power is calculated from measured radio field intensities.

The general symbols used in this chapter are the same as those used in Section 1 (page 1). The following additional symbols are used in this Section:

E_D = radio field intensity of direct wave, in volts per meter

F = radiation efficiency of antenna system

h_1 = elevation above ground of transmitting antenna, in meters

h_2 = elevation above ground of receiving antenna, in meters

P_e = equivalent radiated power, in watts

P_i = power input to antenna, in watts

ϵ = specific inductive capacitance of the ground, i.e., ratio of permittivity (dielectric constant) of the ground to that of air

σ = ground conductivity, in mhos per meter

2.2 EQUIVALENT RADIATED POWER FOR GROUND-WAVE TRANSMISSION

This method is primarily useful at frequencies below 5 megacycles. The equivalent radiated power for ground-wave transmission may best be determined by the measurement of radio field intensity near the antenna. The method applies for the antenna at the ground or not more than 1 wavelength above the ground. If the soil were a perfect conductor; if the radiation pattern of the antenna were perfectly circular; if all measuring points gave absolutely reliable results; and if the distance of the measuring point from the antenna were determined with perfect accuracy; the radiated power could be obtained by the measurement of the radio field intensity at a single point and by using the following general relation between radiated power and measured radio field intensity:

$$P_e = \frac{E_1^2}{E_2^2} \tag{19}$$

where E_1 = measured inverse-distance radio field intensity at 1 kilometer

E_2 = radio field intensity from an ideal antenna having the same effective length as the actual antenna, at 1 kilometer, for 1 watt radiated, from Fig. 9.

The relation between radio field intensity and effective length[12,13] shown by Fig. 9 assumes an ideal antenna having a sinusoidal current distribution.

Actually the theoretical measuring conditions enumerated above cannot be expected in practice, and the accurate determination of the 1-kilometer radio field intensity is considerably more involved than the making of a single measurement of radio field intensity. Possible errors in measuring the radio field intensity near the antenna include (a) errors in determining the exact distance of the measuring point from the antenna; (b) the possibility that the value of radio field intensity at that particular measuring point is not representative but is influenced by directivity of the radiated pattern, uneven terrain, proximity of overhead wires, and in some cases underground conductors; and (c) inherent inaccuracy of the radio field-intensity meter plus the inaccuracy caused by the observer in the operation of the equipment.

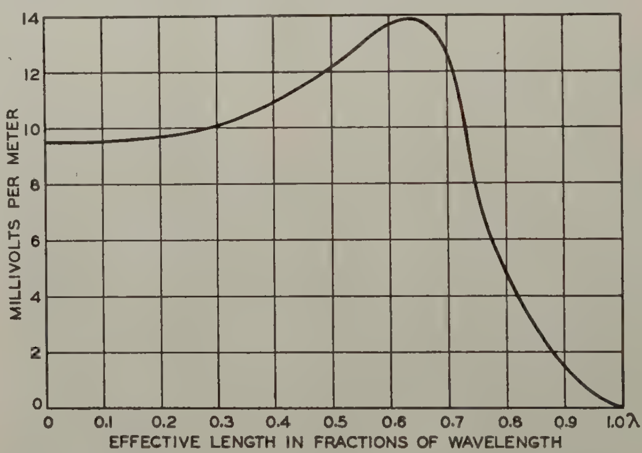


Fig. 9—Unattenuated radio field intensity at 1 kilometer from an ideal antenna, for 1 watt radiated.

The measurements of radio field intensity should be made on a number of equally spaced radials, each extending from about 0.8 kilometer (about ½ mile) to about 25 kilometers (about 15 miles) from the antenna. These distances may be reduced, for frequencies above 1600 kilocycles, approximately in the square of the ratio of 1600 to the frequency in kilocycles. The number of radials, the distance to which each radial is carried, and the number of measurements to be made on each radial, will depend inversely upon how good the measurement points are. In general, 8 radials should be used and 30 to 40 measurements should be made on each radial. An approximate guide which may be followed is to make the distance interval between successive points along the radials approximately equal to 10 per cent of the distance of these

[12] Stuart Ballantine, "On the optimum transmitting wave length for a vertical antenna over perfect earth," Proc. I.R.E., vol. 12, pp. 833–840; December, 1924.

[13] Stuart Ballantine, "High quality radio broadcasting," Proc. I.R.E., vol. 22, pp. 564–629; May, 1934.

points to the transmitting antenna. The data corresponding to each radial should be plotted on log-log co-ordinate paper similar to that of Fig. 8 with the values of radio field intensity as ordinates and distances as abscissas. Radio field intensity and distance may be in any units. It is preferable to use thin, almost transparent, co-ordinate paper unless a light-table is available.

The data should be plotted only for distances greater than 1 wavelength (or, for single antennas higher than 0.2 wavelength, five times the vertical height of the antenna; or for directional antennas, ten times the spacing between elements). The graph paper plotting the measured data is superimposed upon the curves of Fig. 8, shifting the data sheet vertically and horizontally (keeping the vertical lines on both sheets parallel) until the plotted data fit best one of the curves of Fig. 8. If it is found desirable to reconstruct Fig. 8 with a different scale of co-ordinates, the data given in the Appendix, page 13, may be used.

On account of the irregularities of data obtained in practice it may be difficult to determine which of the curves is best fitted by the plotted data. In such case it is desirable to aid the choice by calculation based on whatever may be known or estimated regarding the ground conductivity and the specific inductive capacitance of the terrain over which the measurements are made. One calculates the angle whose tangent is $1/1.8(10)^{10}\ (\epsilon f/\sigma)$. In fitting the plotted data to the curves, one then gives preference to that curve marked with the angle nearest this value.

When the data sheet is finally adjusted, the value of radio field intensity corresponding to the intersection of the inverse-distance line of Fig. 8 with the 1-kilometer ordinate on the data sheet is the inverse-distance radio field intensity at 1 kilometer. (The value of radio field intensity corresponding to the intersection of the curve of best fit of Fig. 8 with the 1-kilometer ordinate on the data sheet is the attenuated radio field intensity at 1 kilometer.)

When the data for all the radials have been analyzed in this manner, a curve should be drawn on polar co-ordinate paper from the inverse-distance 1-kilometer radio field intensities obtained, which gives the unabsorbed field pattern at 1 kilometer. The radius of a circle, the area of which is equal to the area bounded by this pattern, is the effective unabsorbed radio field intensity at 1 kilometer. This value of radio field intensity corresponds to E_1 in (19). E_2 is next determined from Fig. 9 based on the assumed effective length of the actual antenna. The equivalent radiated power in watts P_e may now be computed from (19).

The radiation efficiency of the antenna system in per cent is

$$F = 100 \frac{P_e}{P_i}. \tag{20}$$

While making the radio field-intensity survey, the output power of the station should be maintained constant.

2.3 EQUIVALENT RADIATED POWER FOR IONOSPHERIC TRANSMISSION

This method is primarily useful at frequencies above 3 megacycles. The radiation from antennas at these frequencies is usually greater at some angle to the horizon than it is along the horizon. For this reason measurements made along the ground are not usually indicative of the main power which is used for transmission.

One of the methods used to measure the equivalent radiated power from a high-frequency antenna is to compare the radiation at a distant point after reflection from the ionosphere with the radiation from an antenna of known characteristics such as a half-wave dipole antenna. This can be done by shifting from one antenna to the other and recording the radio field intensity continuously. This should be done a sufficient number of times, sufficiently spaced, to average out the errors caused by fading. The comparison antenna should be placed at the same height from the ground as the height of the center of the antenna being tested.

A refinement of this method is to measure the difference in gain of the two transmitting antennas by connecting the two antennas to two transmitters operating on the same frequency and keyed alternately by fast dots,[14] e.g., 100 words per minute, of unequal lengths. The resulting field is observed at the receiving station by means of a cathode-ray oscillograph. During the observations the power of the transmitter supplying the antenna of greater gain is reduced until the two fields are equal at the receiving station. Then the relative power supplied to the two antennas is the measure of gain.

The equivalent radiated power from a half-wave dipole antenna in free space, in watts, is

$$P_e = 70 \cdot I^2, \tag{21}$$

where I is the current measured at center of dipole antenna, in amperes. When the dipole antenna is less than approximately 1 wavelength from surrounding objects the radiation resistance is slightly different and

[14] Andrew Alford, "A method for measuring the gain of transmitting antennas," Presented, Joint Meeting, U.R.S.I.-I.R.E., Washington, D. C., April 28, 1939. Not published. Abstract:

"Difference in gain of two transmitting antennas may be accurately measured by connecting the two antennas to two transmitters operating on the same frequency and keyed alternately by fast dots, e.g., 100 words per minute, of unequal lengths. The resulting signal is observed at the receiving station by means of a cathode-ray oscillograph. During the observations the power of the transmitter supplying the antenna of greater gain is reduced until the two signals are equal at the receiving station. Then the relative power supplied to the two antennas is the measure of gain."

a correction factor must be used[15–17] with (21). When using a dipole antenna to determine the gain of a directive antenna, considerable care must be exercised to eliminate all stray coupling between the directive antenna and the comparison antenna.[18]

It is possible to measure the equivalent radiated power of a high-frequency antenna by measuring the direct-wave field using an airplane or balloon to support the receiving antenna. If an airplane is used, care must be taken to make sure that the receiving-antenna pattern is not distorted by the presence of the airplane, or, if it is distorted, this must be taken into consideration. The direct-wave radio field intensity[19] from a half-wave dipole antenna suspended in free space, in volts per meter, is

$$E_D = \frac{7.0\sqrt{P_e}}{d}. \tag{22}$$

[15] P. S. Carter, "Circuit relations in radiating systems and applications to antenna problems," Proc. I.R.E., vol. 20, pp. 1004–1041; June, 1932.

[16] H. T. Friis, C. B. Feldman, and W. M. Sharpless, "The determination of the direction of arrival of short radio waves," Proc. I.R.E., vol. 22, pp. 47–78; January, 1934.

[17] C. R. Burrows, "Radio propagation over plane earth,—Field strength curves," *Bell Sys. Tech. Jour.*, vol. 16, pp. 45–77 and 574–577; June and October, 1937.

[18] E. J. Sterba, "Theoretical and practical aspects of directional transmitting systems," Proc. I.R.E., vol. 19, pp. 1184–1215; July, 1931.

[19] B. Trevor and P. S. Carter, "Notes on propagation of waves below ten meters in length," Proc. I.R.E., vol. 21, pp. 387–426; March, 1933.

2.4 EQUIVALENT RADIATED POWER FOR ULTRA-HIGH FREQUENCIES

These methods are for frequencies over 30 megacycles. The equivalent radiated power may be calculated from (22), if the direct-wave field is known. The measurement of the direct-wave field is ordinarily difficult on the ground, since the field at any point is the sum of the fields of the direct wave and the ground-reflected wave. Two methods have been used on ultra-high frequencies to evaluate the direct-wave field. They may be called the variable-height method and the variable-distance method.

With the first method,[20] the receiving antenna is set up about 10 or 12 wavelengths away from the transmitting antenna. The receiving-antenna height is changed and, hence, the phase relations between the direct and ground-reflected waves varied, so that both the sum and difference of these two components are obtained. The direct-wave component is then

$$E_D = \frac{E_3 + E_4}{2}, \tag{23}$$

where E_D = direct-wave radio field intensity,

E_3 = measured radio field intensity when the height of the receiving antenna is adjusted for maximum field intensity,

E_4 = measured radio field intensity when the height of receiving antenna is adjusted for minimum field intensity.

Care should be taken when using this method to make the radio field-intensity measurements over fairly level terrain, since an uneven terrain will introduce a possibility of error.

The second method of measuring the direct-wave field, called the variable-distance method, has been used on ultra-high frequencies below 300 megacycles (1 meter).[21] The procedure is to move the receiving antenna away from the transmitting antenna, measuring the maximum and minimum values of the received field E_3 and E_4, respectively, and recording the respective distances d_3 and d_4. The receiving and transmitting antennas should be kept several wavelengths above the ground and the data carried out to about 50 wavelengths from the transmitter. Then the direct-wave radio field intensity at an intermediate distance d is

$$E_D = \frac{E_3 d_3 + E_4 d_4}{2d}. \tag{24}$$

The equivalent radiated power can then be calculated using (22).

The comparison method of measuring equivalent radiated power described in Section 2.2 may also be used for ultra-high frequencies. The principal application of this method is to compare directive or special (wide-band etc.) antennas with simple antennas. These comparisons should be made somewhat inside the horizon so as to minimize possible errors due to fading.

Another method of measuring the equivalent radiated power on ultra-high frequencies is to take a large number of measurements of radio field intensity on a radial from the transmitter out to 30 or 40 kilometers (about 20 or 25 miles). It is preferable to make continuous recordings using mobile recording equipment.[22] The measurements should be made over level ground free from major obstructions. The recording of radio

[20] C. R. Burrows, A. Decino, and L. E. Hunt, "Ultra-short-wave propagation over land," Proc. I.R.E., vol. 23, pp. 1507–1535; December, 1935.

[21] B. Trevor and R. W. George, "Notes on propagation at a wavelength of seventy-three centimeters," Proc. I.R.E., vol. 23, pp. 461–469; May, 1935.

[22] G. S. Wickizer, "Field-strength survey 52.75 megacycles from Empire State Building," Proc. I.R.E., vol. 28, pp. 291–296; July, 1940.

field intensity will, of course, show considerable variations caused by multiple reflections. By averaging the data it is possible to arrive at the following relation:

The received radio field intensity in volts per meter from a half-wave dipole antenna at grazing angles and at distances within the horizon may be shown to be[23]

$$E = 2.93(10)^{-7} \frac{\sqrt{P_e}\, f h_1 h_2}{d^2}, \tag{25}$$

or

$$E = \frac{88\sqrt{P_e}\, h_1 h_2}{\lambda d^2}, \tag{26}$$

where h_1 = elevation above ground of the transmitting antenna, in meters, and
h_2 = elevation above ground of the receiving antenna, in meters.

[23] H. H. Beverage, "Some notes on ultra-high-frequency propagation," *RCA Rev.*, vol. 1, pp. 76–87; January, 1937.

From (25) or (26) it will be noted that for the conditions assumed, the radio field intensity is inversely proportional to the square of the distance.

The averaged data obtained from the mobile recordings are plotted on logarithmic co-ordinate paper with radio field intensities as ordinates and distance as absciassas. A straight line, having an inverse-square slope is drawn through the averaged data. From any point on this line the equivalent radiated power may be calculated using (25) or (26).

Substituting the value of P_e measured by one of the other methods, the equivalent elevations of the antennas above ground can be calculated from this relation. If the receiving locations are free from interfering objects, the equivalent elevation h_2 is the actual elevation, and the equation permits calculating h_1.

On account of possible effects from buildings and other objects near the antennas, the method must be used with caution.

2.5 ADDITIONAL REFERENCE

J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Company, New York, N. Y., 1941.

2.6 APPENDIX

Users of this process of measuring the power radiated from an antenna may prefer to make their own copies of Fig. 8 on co-ordinate paper of their own choosing and thus have the figure on paper of the exact size as that on which they plot their measurement data. Fig. 8 itself has co-ordinates that match approximately those of Keuffel and Esser log-log co-ordinate paper No. 358-127 (translucent sheet) and 358-127G (opaque sheet). Table I of the data from which Fig. 8 was constructed will permit anyone to draw the curves on any paper desired. Data are not given for the 180-degree curve, as that is not used in cases which occur in actual practice.

TABLE I

Distance	Inverse Distance	Radio Field Intensity																		
		0 degrees	5 degrees	10 degrees	15 degrees	20 degrees	25 degrees	30 degrees	35 degrees	40 degrees	45 degrees	50 degrees	55 degrees	60 degrees	65 degrees	70 degrees	75 degrees	80 degrees	85 degrees	90 degrees
0.01	100.	99.57	98.80	98.04	97.30	96.56	95.85	95.17	94.48	93.82	93.17	92.54	91.92	91.33	90.76	90.21	89.78	89.17	88.68	88.24
0.0121	—	82.22	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.0144	—	69.02	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.0169	—	58.74	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.0196	—	50.59	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.02	—	49.57	49.05	48.50	48.00	47.45	46.95	46.50	46.00	45.55	45.15	44.70	44.30	43.90	43.55	43.20	42.87	42.54	42.21	41.89
0.0225	—	44.02	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.0256	—	38.64	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.0289	—	34.18	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.0324	—	30.44	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.0361	—	27.28	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.04	—	24.57	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	19.47
0.0441	—	22.25	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.0484	—	20.24	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.05	—	19.58	19.24	18.92	18.60	18.30	18.02	17.74	17.46	17.20	16.94	16.70	16.47	16.24	16.03	15.82	15.65	15.49	15.32	15.15
0.0529	—	18.48	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.0576	—	16.94	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.06	—	16.24	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	12.27
0.0625	—	15.58	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.0676	—	14.37	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.09	—	10.69	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.10	—	9.581	9.35	9.13	8.92	8.72	8.53	8.34	8.16	7.99	7.840	7.680	7.545	7.410	7.285	7.160	7.055	6.950	6.845	6.740
0.16	—	5.836	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.24	—	3.754	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	2.270
0.25	—	3.595	3.464	3.34	3.228	3.120	3.02	2.924	2.832	2.752	2.676	2.600	2.534	2.468	2.412	2.356	2.305	2.254	2.204	2.153
0.26	—	3.442	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	2.045
0.36	—	2.382	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.49	—	1.657	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.50	—	1.617	1.538	1.466	1.398	1.336	1.28	1.226	1.178	1.132	1.094	1.054	1.021	0.988	0.959	0.930	0.9075	0.885	0.8625	0.84
0.64	—	1.185	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.74	—	0.9878	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	0.4751
0.75	—	0.9693	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	0.466
0.76	—	0.9539	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	0.4566
0.81	—	0.8767	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
1.00	—	0.6565	0.616	0.579	0.546	0.516	0.489	0.465	0.443	0.423	0.406	0.389	0.375	0.361	0.3495	0.338	0.3286	0.3192	0.397	0.3003
1.21	—	0.4978	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
1.44	—	0.3811	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
1.50	—	0.3567	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	0.1571
1.69	—	0.2939	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
1.96	—	0.2279	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
2.00	—	0.2199	0.204	0.1905	0.1785	0.1680	0.1585	0.1505	0.143	0.136	0.1305	0.125	0.1205	0.116	0.1122	0.1085	0.1056	0.1027	0.09976	0.09685
2.25	—	0.1776	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
2.50	—	0.1454	0.1352	0.1264	0.1188	0.1120	0.1060	0.1004	0.0960	0.0916	0.088	0.0844	0.0814	0.0784	0.076	0.0736	0.07164	0.06968	0.06772	0.06576
2.56	—	0.1389	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
2.89	—	0.1090	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
3.00	—	0.1009	0.094	0.0880	0.0830	0.07867	0.0730	0.0710	0.0680	0.0650	0.06267	0.0600	0.0580	0.0560	0.05433	0.05267	0.05139	0.05012	0.04884	0.04757
3.24	—	0.08581	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
3.61	—	0.06783	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
4.00	—	0.05384	0.05075	0.048	0.0455	0.04350	0.0415	0.03975	0.03825	0.0370	0.03575	0.03475	0.03375	0.03275	0.03188	0.031	0.03029	0.02958	0.02886	0.02815
5.00	—	0.03186	0.0304	0.0292	0.0280	0.02700	0.0262	0.02540	0.02460	0.0238	0.0232	0.0226	0.02210	0.0216	0.0211	0.0206	0.02009	0.01958	0.01907	0.01856
6.00	—	0.02047	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
7.00	—	0.01406	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
8.00	—	0.01020	0.01005	0.009888	0.009725	0.009563	0.009363	0.009225	0.009063	0.00890	0.008738	0.008575	0.008425	0.008275	0.008138	0.008	0.007886	0.007772	0.007659	0.007545
9.00	—	0.007733	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	0.0060011
10.00	—	0.006060	0.00604	0.00602	0.00599	0.00595	0.00588	0.00577	0.00567	0.00556	0.00545	0.00535	0.00525	0.00515	0.005075	0.0050	0.004971	0.004942	0.004913	0.004884
11.00	—	0.004900	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
12.00	—	0.004042	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
13.00	—	0.003392	0.003392	0.003392	0.003385	0.003377	0.003369	0.003354	0.003331	0.003308	0.003269	0.003231	0.003181	0.003131	0.003073	0.003015	0.00299	0.002965	0.002939	0.002914
15.00	—	0.002493	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
18.00	—	0.001694	0.001694	0.001694	0.001694	0.001694	0.001694	0.001694	0.001689	0.001689	0.001678	0.001672	0.001656	0.001639	0.001619	0.0016	0.001582	0.001565	0.001547	0.001529
20.00	—	0.001360	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	0.001242
30.00	—	0.0005867	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
40.00	—	0.0003250	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
50.00	—	0.0002064	0.000206	0.000206	0.000206	0.000206	0.000206	0.000206	0.000206	0.000206	0.000206	0.000206	0.000206	0.000206	0.000206	0.000206	0.000206	0.000206	0.000206	0.000206
100.0	—	0.000050	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
200.00	0.005	0.0000125	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	0.0000125

SECTION 3. METHODS OF MEASURING NOISE FIELD INTENSITY

3.1 GENERAL

The measurement of noise field intensity, while involving the same basic procedure as the measurement of radio field intensity, presents additional problems because of the fact that, in general, the direction of propagation, polarization, relative phases, and amplitudes of the components of noise fields are not well defined and are subject to variations with reference to time and space.

The measuring apparatus, including the antenna, is identical with that employed in measuring radio field intensity, except that the output circuit must satisfy special requirements as to time constant, and the effective bandwidth and phase characteristics of the selective circuits must be taken into account.

The measuring apparatus is tuned to the frequency at which the noise field intensity is to be measured, and the voltage produced in the receiving antenna by the noise is measured. Since the voltmeter is usually calibrated in terms of a known voltage of single frequency, the noise field intensity is measured in terms of the equivalent effect of a single-frequency field and of the effective bandwidth of the radio receiver. The time constants of the output circuit and the nature of the rectifier determine whether the average, root-mean-square, peak, or quasi-peak value of the noise field intensity is measured. Since the noise field is generally distributed over a wide frequency range, the noise field intensity measured is a function of the effective bandwidth of the radio receiver; hence, accurate specification of the effective bandwidth of the measuring set is important.

The general method of determining the effective bandwidth of the measuring set, and the manner in which it influences the measured value of noise field intensity, are indicated in the definitions of effective bandwidth and noise intensity.

It is also of importance to record a classification of the noise (e.g., ignition noise, atmospherics, random noise, impulse noise), since there may be a difference in the radio-field-to-noise ratio required for operation through various types of noise.

3.2 MEASUREMENT METHODS

The measurement of particular characteristics of noise field intensity which bear a definite relationship to the disturbing effects of the noise would, if found feasible, provide means for establishing the degree of disturbance which the noise causes to the different types of communication. Specialized methods in this category have been developed for measurement of the disturbance caused by noise to telephone (voice and music) signals. This apparatus takes the form of a field-intensity-measuring set of the standard-antenna type with a special output circuit which includes a rectifier, an integrating circuit, and an indicating instrument. From experience gained with such measurements, it has been determined that the effect of noise in telephony (voice or music) is more nearly proportional to the peak intensity than to the average intensity. A joint committee of the Edison Electric Institute, National Electrical Manufacturers Association, and Radio Manufacturers Association accordingly recommended[24] the measurement of a quasi-peak value, the specified time constants of the output indicator circuit being 10 milliseconds charge and 600 milliseconds discharge; an effective bandwidth of 10 kilocycles was recommended.

A single quasi-peak value is not in general proportional to the disturbing effect of all types of radio noise and for all types of signals. Thus, for wide-band service requirements, it may be desirable to make noise measurements with apparatus of greater effective bandwidth and indicating devices with the equivalent of lower time constants.

A cathode-ray oscillograph may be advantageously used in measuring noise peaks. It has the advantage of showing the noise wave shape. It has the disadvantage, however, that if the noise peak is very short the top of the peak is hard to read, unless the room is darkened, because of the low illumination for a transient pulse.

3.21 ATMOSPHERICS

Both short-time and long-time distributions of the noise field intensity of atmospherics are of interest. Measurements by automatic-recorder methods of the time distribution of both the average and peak values provide data in a desirable form for most purposes.

For recorder operation, it is possible to measure the average intensity by using approximately equal charge and discharge time constants of the order of 1 minute.

It is possible to measure the instantaneous peak intensity in terms of a quasi-peak value of very short duration, of the order of 0.1 millisecond; for recorder operation, a discharge time constant of the order of 1 minute has been found desirable.

The average or quasi-peak data (or both) may be presented in the form of time distribution curves, as in recording received intensities from radio stations which are fading; and provide information on the noise field intensity of atmospherics at a given place for a given percentage of time of the period considered.

[24] C. V. Aggers, D. E. Foster, and C. S. Young, "Instruments and methods of measuring radio noise," *Trans. A.I.E.E.* (*Elec. Eng.*, March, 1940), vol. 59, pp. 178–192; March, 1940.

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Standards

on

RADIO WAVE PROPAGATION

DEFINITIONS OF TERMS

1942

Supplement to the Proceedings of the I. R. E.
Vol. 30, No. 7, Part III

THE INSTITUTE OF RADIO ENGINEERS

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RADIO WAVE PROPAGATION

DEFINITIONS OF TERMS

1942

The Institute of Radio Engineers, Inc.
330 West 42nd Street
New York, N. Y.

CONTENTS

INTRODUCTION

These definitions of terms are the result of more than two years of study by the Technical Committee on Radio Wave Propagation, carried on in frequent meetings and by correspondence under the general guidance of the Standards Committee. Published with the approval of the Board of Directors, the report embodies the Institute's official recommendations to its members and the industry at large.

Suggestions and comments will be welcomed as an aid to committees preparing future reports. Correspondence should be addressed to the Institute of Radio Engineers.

CONCERNING THE INSTITUTE AND ITS STANDARDS ACTIVITIES

The Institute appointed its first standards committee in 1912, and the next year published a report dealing with definitions of terms, letter and graphical symbols, and methods of testing and rating equipment. Expanded reports appeared in 1915, 1922, 1926, 1928, 1931, and 1933, each of which combined, in a single document, data on all branches of the art.

Publication of the current series of standards, of which this one is a part, was begun in 1938.* It differs from earlier reports in that each individual booklet deals with a separate field. Under present policies, subdivision is being carried even farther and separate booklets are being issued in each field for definitions of terms, for symbols, and for measuring and testing methods.

Beginning with 1942, all standards are being published in the 8 1/2-×11-inch size to conform with the new format for the PROCEEDINGS of the I.R.E.

Distribution of Standards Reports

The Institute is mailing one copy of this report to every member who is in good standing for 1942.

Additional copies, if available, may be obtained by members or nonmembers from the Institute office at the price mentioned on the inside back cover.

Co-operation with Other Organizations

Throughout its life, the Institute has co-operated with other bodies in the establishment of standards. Last year, for instance, there were more than 50 official I.R.E. delegates to other standardization groups. The Institute is also the sponsor for the American Standards Association's Sectional Committee on Radio.

The Institute of Radio Engineers

The Institute of Radio Engineers was founded in 1912 to advance the theory and practice of radio and allied branches of engineering and of the related arts and sciences, their application to human needs, and the maintenance of a high professional standing among its members. Although mostly located in the United States of America, the Institute membership of over 7000 persons has large representation throughout the world.

The PROCEEDINGS, which has been published since 1913, is issued monthly and contains contributions from the leading workers in the theoretical and practical fields of radio communication. It is forwarded to all members, who receive also the various standards reports which are published at irregular intervals.

Applications for membership are invited from those interested in radio. Full information may be obtained from the Secretary.

* For a detailed list of current standards reports, see the inside back cover.

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SECTION 1. GENERAL

1W1. Radio Wave Propagation. The transfer of energy by electromagnetic radiation at radio frequencies.

1W2. Radio Frequency. A frequency at which electromagnetic radiation of energy is useful for communication purposes.

NOTE: The present useful limits of radio frequencies are roughly 10 kilocycles to 10,000 megacycles.

1W3. Pulse. A transient disturbance.

1W4. Displacement. A change in a medium, proportional to the square-root of the stored energy of a certain kind. It is exemplified by compression in a sound wave, and by electric or magnetic flux density in an electromagnetic wave.

1W5. Electric Displacement. Electric flux density or electric induction.

1W6. Magnetic Displacement. Magnetic flux density or magnetic induction.

1W7. Wave. A disturbance propagated through a medium. Also, the graphical representation of a wave or of any periodic variation.

1W8. Sinusoidal Wave. A wave whose displacement is the sine (or cosine) of an angle proportional to time or distance or both.

1W9. Periodic Wave. A wave whose displacement has a periodic variation with time or distance or both.

1W10. Traveling Wave. A wave traveling in one direction.

1W11. Stationary Wave. A periodic wave which at all points in space has simultaneous proportionate variations with time.

1W12. Wave Interference. The variation of wave amplitude with distance or time, caused by variations of relative phase in the combination of two or more waves.

1W13. Envelope Delay. The time of delay of the envelope of a wave. It is equal to the slope of phase angle against angular frequency and is defined only for a frequency band over which the slope is uniform.

1W14. Group Velocity. The velocity of propagation of the envelope of a wave occupying a frequency band over which the envelope delay is uniform. It differs from phase velocity only in a medium in which the phase velocity varies with frequency or direction.

1W15. Phase Velocity. Of a periodic wave, the velocity with which a point of a certain phase travels in the direction of propagation.

1W16. Wave Front. At the beginning of a wave, the surface over which the displacement from zero starts at the same instant. In a continuous wave, the surface over which the displacements at the same instant are in the same phase.

1W17. Wave Normal. The direction normal to the wave front and toward the direction of propagation.

1W18. Wavelength. In a periodic wave, the distance between corresponding phases of two consecutive cycles. It is equal to the quotient of phase velocity by frequency.

1W19. Radianlength. In a sinusoidal wave, the distance between phases differing by an angle of one radian. It is equal to the wavelength divided by 2π.

1W20. Angular Frequency (Radian Frequency). The frequency expressed in radians per unit of time. It is equal to the frequency in cycles multiplied by 2π.

1W21. Angular Distance. The distance expressed in radianlengths, or equivalent angular measure. It is equal to 2π radians or 360 degrees, multiplied by the distance in wavelengths.

1W22. Angular Length. The length expressed in radianlengths, or equivalent angular measure. It is equal to 2π radians or 360 degrees, multiplied by the length in wavelengths.

1W23. Incident Wave. A wave in a medium of certain propagation characteristics impinging on a medium of different propagation characteristics.

1W24. Reflected Wave. The wave caused by the reflection of part of an incident wave back into the first medium.

1W25. Refracted Wave. The wave caused by the refraction of the part of an incident wave which travels into the second medium.

1W26. Reflection Coefficient. The ratio of a quantity in the reflected wave to the same quantity in the incident wave. The reflection coefficient may be different for different functions of the wave (field intensity, power, etc.).

1W27. Refractive Index. Of a wave-transmission medium, the ratio of the phase velocity in free space to that in the medium.

1W28. Relative Refractive Index. Of two mediums, the ratio of their refractive indexes.

1W29. Transverse Wave. A wave in which the displacement is transverse to the direction of propagation.

1W30. Electromagnetic Wave. A wave in which there are both electric and magnetic displacements. Electromagnetic waves are known as radio waves, heat rays, light, X rays, etc., depending on the frequency.

1W31. Transverse Electric Wave. A wave of electric displacement transverse to the direction of propagation.

NOTE: This is abbreviated, TE wave.

1W32. Transverse Magnetic Wave. A wave of magnetic displacement transverse to the direction of propagation.

NOTE: This is abbreviated, TM wave.

1W33. Transverse Electromagnetic Wave. An electromagnetic wave in which both electric and magnetic displacements are transverse to the direction of propagation.

NOTE: This is abbreviated, TEM wave.

1W34. Hybrid Electromagnetic Wave. An electromagnetic wave which has both transverse and longitudinal components of displacement.

1W35. Plane Wave. A wave whose wave front is a plane surface.

1W36. Uniform Plane Wave. A plane wave in which the displacement is uniform over the wave front.

1W37. Spherical Wave. A wave whose wave front is a spherical surface.

1W38. Cylindrical Wave. A wave whose wave front is a cylindrical surface.

1W39. Plane-Polarized Wave. A transverse wave in which the direction of the displacement at all points in a certain space is parallel to a fixed plane parallel to the direction of propagation.

1W40. Linearly Polarized Wave. A transverse wave in which the displacement has a constant direction at a point in space.

1W41. Direction of Polarization. In a linearly polarized wave, the direction of the displacement vector. In an electromagnetic wave, the direction of the electric displacement is taken as the direction of polarization.

NOTE: This convention is in accord with the present trend in optics but not with earlier usage in which the direction of the magnetic displacement was taken as the direction of polarization.

1W42. Plane of Polarization. In a plane-polarized wave, the fixed plane parallel to the direction of polarization and the direction of propagation.

1W43. Vertically Polarized Wave. A linearly polarized wave whose direction of polarization is vertical.

1W44. Horizontally Polarized Wave. A linearly polarized wave whose direction of polarization is horizontal.

1W45. Elliptically Polarized Wave. A wave in which the direction of displacement at a point rotates about a point in a plane and the magnitude of displacement varies as the radius of an ellipse.

1W46. Circularly Polarized Wave. An elliptically polarized wave in which the direction of displacement at a point rotates with constant angular velocity about an axis in the direction of the propagation and the magnitude of displacement is independent of its direction.

1W47. Right-Handed Elliptically Polarized Wave. An elliptically polarized wave in which the rotation of the direction of displacement is clockwise for an observer looking in the direction the wave is traveling.

1W48. Left-Handed Elliptically Polarized Wave. An elliptically polarized wave in which the rotation of the direction of displacement is counterclockwise for an observer looking in the direction the wave is traveling.

1W49. Attenuation. Of a wave, the decrease in displacement with distance in the direction of propagation. If the attenuation varies with frequency, it is defined for a sinusoidal wave of a certain frequency and of constant amplitude at any point. The attenuation of a wave may be defined relative to the attenuation in some ideal conditions such as in free space or over a perfectly conducting plane.

1W50. Plane-Earth Attenuation. The attenuation over an imperfectly conducting plane earth in excess of that over a perfectly conducting plane.

1W51. Spherical-Earth Attenuation. The attenuation over an imperfectly conducting spherical earth in excess of that over a perfectly conducting plane.

1W52. Shadow Attenuation. The attenuation over a sphere in excess of that over a plane, the distance over the surface and other factors being the same.

1W53. Propagation Ratio. In a wave from one point to another, the ratio of the displacement at the second point to that at the first point. It is expressed as a vector involving both amplitude and phase angle.

NOTE: This is also called propagation factor, transfer ratio, or transfer factor.

1W54. Attenuation Ratio. The magnitude of the propagation ratio.

1W55. Phase-Propagation Ratio. The propagation ratio divided by its magnitude. It is expressed as a unit vector of the same angle as the propagation ratio.

1W56. Absorption. The loss of energy from a wave by dissipation in propagation through or adjacent to a dissipative medium.

1W57. Absorption Ratio. The ratio of the displacement at a point in a wave to the value it would have if there were no absorption.

1W58. Atmospheric Radio Wave. A radio wave that is propagated by reflection in the atmosphere. It may include either or both of the components, ionospheric wave (definition 1W59) and tropospheric wave (definition 1W60).

1W59. Ionospheric Wave. A radio wave that is propagated by reflection from the ionosphere (definition 3W1).

NOTE: This is sometimes called a sky wave.

1W60. Tropospheric Wave. A radio wave that is propagated by reflection from a place of abrupt change in the dielectric constant or its gradient with position in the troposphere (definition 1W61).

1W61. Troposphere. That part of the earth's atmosphere in which temperature generally decreases with altitude, clouds form, and convection is active.

NOTE: Experiments indicate that the troposphere occupies the space above the earth's surface to a height of about 10 kilometers.

1W62. Ground Wave. A radio wave that is propagated over the earth and is ordinarily affected by the presence of the ground. The ground wave includes all components of a radio wave over the earth except ionospheric waves and tropospheric waves. The ground wave is somewhat refracted by the normal gradient of the dielectric constant of the lower atmosphere.

1W63. Direct Wave. A wave that is propagated directly through space.

1W64. Ground-Reflected Wave. The component of the ground wave that is reflected from the ground.

1W65. Horizon. In radio wave propagation over the earth, the line which bounds that part of the earth's surface reached by the direct wave. On a spherical surface, the horizon is a circle. The distance to the horizon is affected by atmospheric refraction.

1W66. Tangential Wave Path. In radio wave propagation over the earth, a path of propagation of a direct wave, which is tangential to the surface of the earth. The tangential wave path is curved by atmospheric refraction.

1W67. Fading. The variation of radio field intensity caused by changes in the transmission medium.

1W68. Selective Fading. Fading which is different at different frequencies in a frequency band occupied by a modulated wave.

1W69. Guided Wave. A wave whose propagation is concentrated in certain directions within or near boundaries between materials of different properties located in a path between two places.

1W70. Wave Guide. A system of material boundaries capable of guiding waves.

1W71. Wave Duct. A wave guide with tubular boundaries capable of concentrating the propagation of waves within its boundaries.

1W72. Dielectric Wire. A rod or filament of dielectric material.

1W73. Electric Dipole or Doublet. A simple antenna comprising a pair of conductors which is capable of radiating an electromagnetic wave in response to a displacement of electric charge from one conductor to the other. For theoretical purposes, the elementary dipole or doublet is so small that its directive properties are independent of its size and shape.

1W74. Magnetic Dipole or Doublet. A simple loop antenna which is capable of radiating an electromagnetic wave in response to a circulation of electric current in the loop. For theoretical purposes, the elementary dipole is so small that its directive properties are independent of its size and shape. It is the magnetic analog of the electric dipole.

1W75. Effective Length. Of an antenna, the length which, when multiplied by the current at the point of maximum current, will give the same product as the length and uniform current of an elementary electric dipole at the same location giving the same radio field intensity in the direction of maximum radiation.

1W76. Effective Height. The effective length of a grounded antenna.

1W77. Radio Field Intensity. The electric or magnetic field intensity at a given location resulting from the passage of radio waves. It is commonly expressed in terms of the electric field intensity. In the case of a sinusoidal wave, the root-mean-square value is commonly stated. Unless otherwise stated, it is taken in the direction of maximum field intensity.

SECTION 2. NOISE

2W1. Radio Interference. An undesired disturbance in reception, or that which causes the undesired disturbance. Radio interference may thus be a disturbance in the radio transmitter, the transmission medium, or the radio receiver. Some examples of radio interference are background interference in the transmitter, undesired electromagnetic disturbance in the transmission medium as by lightning or undesired radio

waves, and hum or thermal agitation in the receiver.

2W2. Noise. Interference whose energy is distributed over a wide band of frequencies.

2W3. Atmospheric Noise. Noise caused by natural electrical discharges in the atmosphere.

NOTE: It is also called "static."

2W4. Selective Interference. Radio interference whose energy is concentrated in a narrow band of frequencies. Some examples are other radio stations on the same or adjacent frequencies, harmonics of other radio stations, and unshielded diathermy equipment.

2W5. Radio Station Interference. Selective interference caused by the radio waves from a station or stations other than that from which reception is desired.

2W6. Electrical Interference. Interference caused by the operation of electrical apparatus other than radio stations. It may be either selective interference or noise, usually the latter.

2W7. Random Noise. Noise due to the aggregate of a large number of elementary disturbances with random occurrence in time.

2W8. Impulse Noise. Noise due to a disturbance having an abrupt change and short duration or to a succession of nonoverlapping such disturbances.

2W9. Crest Factor. The ratio of the peak value to the effective (root-mean-square) value of a measure proportional to the square root of the power.

2W10. Form Factor. The ratio of the root-mean-square value to the average value of a measure proportional to the square root of the power.

2W11. Noise Field Intensity. The field intensity of noise in a transmission medium. Its value is defined only with reference to a definite frequency band. For random noise, the noise intensity within a limited frequency band is proportional to the square root of the effective bandwidth (definition 2W14). For impulse noise, if the bandwidth is sufficient not to cause overlapping of the impulses, the peak value of noise field intensity is proportional to the bandwidth and the average value is independent of the bandwidth. For either random or impulse noise, the root-mean-square value is proportional to the square root of the bandwidth and the mean-square value is proportional to the bandwidth. Thus, a complete specification of noise intensity may or may not be possible, depending on the character of the noise.

2W12. Radio-Field-to-Noise Ratio. The ratio, at a given location, of the radio field intensity of the desired wave to the noise field intensity.

2W13. Signal-to-Noise Ratio. The ratio, at a given location, of a measure of the signal to the same measure of the total noise.

NOTE: For the definition of "Signal," see 1R21 in "Standards on Radio Receivers, 1938."

2W14. Effective Bandwidth. Of a band-pass filter, the width of a hypothetical "rectangular" band-pass filter which would pass the same mean-square value of noise current or voltage, with the same transfer ratio at a reference frequency (usually mid-band). It is equal to the integrated product of the incremental bandwidths multiplied by the square of (the current or voltage transfer ratio divided by this ratio at the reference frequency). It is therefore equal to the area under a curve plotted on linear co-ordinates, whose ordinates are the square of this quotient and whose abscissas are the frequency.

SECTION 3. IONOSPHERE

3W1. Ionosphere. That part of the earth's atmosphere above the lowest level at which the ionization is large compared with that at the ground, so that it affects the transmission of radio waves.

NOTE: Experiments indicate that this lowest level is about 50 kilometers above the earth's surface.

3W2. Equivalent Electron Density. In an ionized gas, the product of ion density by the mass ratio of an electron to an ion. Where there are several kinds of ions present it is the sum of their individual equivalent electron densities.

3W3. E Region. The region of the ionosphere between about 90 and 140 kilometers above the earth's surface.

3W4. E Layer. An ionized layer in the E region.

3W5. F Region. The region of the ionosphere between about 140 and 400 kilometers above the earth's surface.

3W6. F Layer. An ionized layer in the F region, existing in the night hemisphere and in the weakly illuminated portion of the day hemisphere. Over the intensely illuminated portion of the day hemisphere two layers exist (see F_1 layer and F_2 layer, below).

3W7. F_1 Layer. The lower of the two ionized layers normally existing in the F region in the day hemisphere.

3W8. F_2 Layer. The higher of the two ionized layers normally existing in the F region in the day hemisphere.

3W9. Vertical-Incidence Transmission. The transmission of a radio wave vertically to the ionosphere and back. The transmission is practically the same for

slight departures from the vertical, as when the radio transmitter and receiver are separated a few kilometers.

3W10. Oblique-Incidence Transmission. The transmission of a radio wave obliquely up to the ionosphere and down again.

3W11. Ordinary Wave. The one of the two components into which a radio wave is split in the ionosphere by the earth's magnetic field, which in the lower parts of the ionosphere is left-handed elliptically polarized if the earth's magnetic field has a positive component in the direction of propagation. (This term is used in a different sense in optics.)

NOTE: This wave is designated by the letter symbol O and is sometimes called the O wave.

3W12. Extraordinary Wave. The one of the two components into which a radio wave is split in the ionosphere by the earth's magnetic field, which in the lower parts of the ionosphere is right-handed elliptically polarized if the earth's magnetic field has a positive component in the direction of propagation.

NOTE: This wave is designated by the letter symbol X and is sometimes called the X wave.

3W13. Virtual Height. Of an ionized layer of the ionosphere, the height at which reflection from a definite boundary surface would cause the same time of travel as the actual reflection, for a wave transmitted from the ground to the ionosphere and reflected back. Virtual height depends on the wave components and the frequency: the value usually stated is for the ordinary wave and for the lowest frequency at which reflection occurs.

NOTE: This is sometimes called equivalent height or effective height.

3W14. Penetration Frequency. Of an ionized layer of the ionosphere, the frequency at which the virtual height for a wave component at vertical incidence has a maximum value caused by penetration of the wave through the layer. Except for the occurrence of sporadic and scattered reflections (definitions 3W15 and 3W16), it is the highest frequency of waves reflected from the layer at vertical incidence. The square of the penetration frequency of the ordinary wave is proportional to the maximum equivalent electron density of the layer.

NOTE: This has been called the critical frequency.

3W15. Sporadic Reflections. From an ionized layer of the ionosphere, sharply defined reflections of substantial intensity from the layer at frequencies greater than the critical frequency of the layer. The intensity of the sporadic reflections generally decreases with increasing frequency. They are variable in respect to time of occurrence, geographic distribution, and frequency range.

NOTE: Sporadic reflections are sometimes called abnormal reflections.

3W16. Scattered Reflections. From a region of the ionosphere, reflections composed of many components of different virtual heights of reflection, which interfere and cause rapid fading. They are variable in respect to time of occurrence, geographical distribution, intensity, and frequency range.

3W17. Zigzag Reflections. From a layer of the ionosphere, high-order multiple reflections which may be of abnormal intensity. They occur in waves which travel by multihop ionospheric reflection and finally turn back toward their starting point by repeated reflections from a slightly curved or sloping portion of an ionized layer.

3W18. Ionospheric Storm. A period of disturbance in the ionosphere, during which there are anomalous variations of penetration frequencies, virtual heights, and absorption. The disturbance usually includes: turbulence of the ionosphere and poorly defined penetration frequencies, especially at night; great virtual heights and low penetration frequencies, especially of the F and F_2 layers; great absorption, especially at night at frequencies of the order of one megacycle. An ionospheric storm usually lasts a substantial part of a day or more than one day.

3W19. Sudden Ionospheric Disturbance. A sudden increase of ionization density in low parts of the ionosphere, caused by a bright solar chromospheric eruption. It gives rise to a sudden increase of absorption in radio waves propagated through the low parts of the ionosphere, and sometimes to simultaneous disturbances of terrestrial magnetism and earth currents. The change takes place within one or a few minutes, and conditions usually return to normal within one or a few hours.

3W20. Radio Fadeout. A cessation or near cessation of propagation of radio waves through the parts of the ionosphere affected by a sudden ionospheric disturbance.

3W21. Line of Propagation. Of a wave from one point to another, the line between these two points which is in the direction of the wave normal at all points in space. This is also called the path or line of travel.

3W22. Wave Angle. Of a wave from one point to another, the angle of the line of propagation. It has two components, azimuth and elevation. The azimuth angle is measured about a vertical axis, clockwise from north. The elevation angle is measured about the horizontal axis to the direction of propagation, upward from horizontal.

3W23. Angle of Departure. Of a wave from transmitter to receiver, the wave angle of the line of propagation, on departing from the transmitting antenna.

3W24. Angle of Arrival. Of a wave from transmitter

to receiver, the wave angle of the line of propagation, on arriving at the receiving antenna.

3W25. Horizontal Angle of Deviation. Of a wave from transmitter to receiver, the horizontal angle between the great-circle path and the direction of departure or arrival along the line of propagation.

3W26. Angle of Incidence. Of a wave arriving at a surface, the angle between the normal to the surface at the point of incidence and the line of propagation on approaching the surface.

3W27. Brewster Angle. The angle of incidence at which a wave polarized in the plane of the angle of incidence undergoes a phase shift of one quadrant on reflection at the surface.

3W28. Hop. An excursion of a radio wave from the earth to the ionosphere and back to earth, in traveling from one point to another. It is usually used in expressions such as single-hop, double-hop, and multihop. The number of hops is called the order of reflection.

3W29. Maximum Usable Frequency. The highest frequency that can be used for radio transmission at a specified time between two points on the earth by reflection from the regular ionized layers of the ionosphere. Higher frequencies are transmitted only by sporadic and scattered reflections.

3W30. Skip Distance. The minimum distance at which radio waves of a specified frequency can be transmitted at a specified time between two points on the earth by reflection from the regular ionized layers of the ionosphere. Reflected waves are received at less distance only by sporadic, scattered, or zigzag reflections.

3W31. Primary Skip Zone. The area around a radio transmitter beyond the ground-wave range but within the skip distance. Radio reception is possible in the primary skip zone only by sporadic, scattered, and zigzag reflections.

3W32. Recombination Coefficient. In an ionized gas, the quotient of the time rate of recombination of ions divided by the product of the positive-ion density and negative-ion density.

3W33. Gyro Frequency. The gyro frequency is the natural frequency of rotation of ions around the lines of magnetic field of the earth. It is equal to the product of 35.38 kilocycles by the earth's magnetic field intensity (ampere-turns per meter) by the mass ratio of the electron to the ion in question. For electrons, it is of the order of 700 to 1600 kilocycles, while for material ions it is in the audio-frequency range.

NOTE: One half this frequency is sometimes called the Larmor precession frequency.

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Vol. 30, No. 7, Part IV

THE INSTITUTE OF RADIO ENGINEERS

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1942

The Institute of Radio Engineers, Inc.
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New York, N. Y.

CONTENTS

INTRODUCTION

These definitions of terms are the result of study by the 1940 Technical Committee on Facsimile, carried on in frequent meetings and by correspondence under the general guidance of the Standards Committee. Published with the approval of the Board of Directors, the report embodies the Institute's official recommendations to its members and the industry at large.

Suggestions and comments will be welcomed as an aid to committees preparing future reports. Correspondence should be addressed to the Institute of Radio Engineers.

CONCERNING THE INSTITUTE AND ITS STANDARDS ACTIVITIES

The Institute appointed its first standards committee in 1912, and the next year published a report dealing with definitions of terms, letter and graphical symbols, and methods of testing and rating equipment. Expanded reports appeared in 1915, 1922, 1926, 1928, 1931, and 1933, each of which combined, in a single document, data on all branches of the art.

Publication of the current series of standards, of which this one is a part, was begun in 1938.* It differs from earlier reports in that each individual booklet deals with a separate field. Under present policies, subdivision is being carried even farther and separate booklets are being issued in each field for definitions of terms, for symbols, and for measuring and testing methods.

Beginning with 1942, all standards are being published in the 8 1/2- ×11-inch size to conform with the new format for the PROCEEDINGS of the I.R.E.

Distribution of Standards Reports

The Institute is mailing one copy of this report to every member who is in good standing for 1942.

Additional copies, if available, may be obtained by members or nonmembers from the Institute office at the price mentioned on the inside back cover.

Co-operation with Other Organizations

Throughout its life, the Institute has co-operated with other bodies in the establishment of standards. Last year, for instance, there were more than 50 official I.R.E. delegates to other standardization groups. The Institute is also the sponsor for the American Standards Association's Sectional Committee on Radio.

The Institute of Radio Engineers

The Institute of Radio Engineers was founded in 1912 to advance the theory and practice of radio and allied branches of engineering and of the related arts and sciences, their application to human needs, and the maintenance of a high professional standing among its members. Although mostly located in the United States of America, the Institute membership of over 7000 persons has large representation throughout the world.

The PROCEEDINGS, which has been published since 1913, is issued monthly and contains contributions from the leading workers in the theoretical and practical fields of radio communication. It is forwarded to all members, who receive also the various standards reports which are published at irregular intervals.

Applications for membership are invited from those interested in radio. Full information may be obtained from the Secretary.

* For a detailed list of current standards reports, see the inside back cover.

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SECTION 1. TERMINAL EQUIPMENT

1F1. Facsimile Transmission. The transmission of signal waves produced by the scanning of fixed graphic material, including pictures, for reproduction in record form.

1F2. Subject Copy. The material in graphic form which is to be transmitted for facsimile reproduction by the recorder.

1F3. Facsimile Transmitter. The apparatus employed to translate the subject copy into signals suitable for delivery to the communication system.

1F4. Scanning. The process of analyzing successively the densities of the subject copy according to the elements of a predetermined pattern or of synthesizing corresponding densities in the recording.

1F5. Rectilinear Scanning. The process of scanning an area in a predetermined sequence of narrow straight parallel strips.

1F6. Scanner. That part of the facsimile transmitter which systematically translates the densities of the elemental areas of the subject copy into signal-wave form.

NOTE: The device generally includes: (1) a light source, (2) an optical system which delineates an elemental area of the subject copy, (3) means for systematically moving one with respect to the other in two dimensions, (4) a light-sensitive device and directly associated circuits.

1F7. Scanning Spot. The area viewed instantaneously by the pickup system of the scanner.

NOTE: The width is measured in the direction of line scanning; height, across this direction.

1F8. Elemental Area. Any segment of a scanning line, the dimension of which along the line is exactly equal to the nominal line width.

1F9. Critical Area. An area of the subject copy whose dimensions along and across the direction of scanning are equal to the definition in these respective directions.

1F10. Density. A measure of the light-reflecting or transmitting properties of an area. It is expressed by the logarithm (base 10) of the ratio of incident to transmitted or reflected light.

1F11. Halftone Characteristic. A relation, usually shown by a graph, between the density of the recorded copy and the density of the subject copy.

NOTE: The term may also be used to relate the amplitude of the facsimile signal to the density of the subject copy or the record copy when only a portion of the system is under consideration. In a frequency-modulation system an appropriate parameter is to be used instead of the amplitude.

1F12. Definition (in a given direction). The width of the narrowest isolated line of the subject copy perpendicular to the given direction, for which the response of the system (or portion thereof under consideration) will just reach the steady-state value which is attained for a larger area of the same density.

1F13. Detail. The square root of the ratio between the number of scanning lines per unit length and the definition in the direction of the scanning line.

NOTE: a) It is essential that the unit length be stipulated.
b) For accurate comparisons this assumes that the definition in the two directions is not widely different.

1F14. Length of Scanning Line. The length of the path traced by the scanning or recording spot in moving from a point on one line to a corresponding point on the next following line.

1F15. Available Line. The portion of the length of the scanning line which can be used specifically for picture signals.

NOTE: The available line may be expressed as a percentage of the length of scanning line.

1F16. Index of Co-operation. In rectilinear scanning, the product of the total length of a scanning line by the number of scanning lines per unit length.

NOTE: This definition should not be confused with the international index of co-operation based on drum diameter and defined by the International Radio Consulting Committee (CCIR). The latter is $1/\pi$ times the former.

1F17. Spot Speed. The product of the length of scanning line by the number of scanning lines per second.

1F18. Drum (or Stroke) Speed. The number of scanning lines per minute.

1F19. Spot Projection. The optical method in which the scanning (or recording) spot is delineated by an aperture between the light source and the subject copy or record sheet.

1F20. Flood Projection. The optical method in which the subject copy is illuminated and the scanning spot is delineated by an aperture between the subject copy and the light-sensitive device.

1F21. Facsimile Receiver. The apparatus employed to translate the signal from the communication channel into a facsimile record of the subject copy.

1F22. Recording. The process of registering the received signal upon the record sheet.

1F23. Photographic Recording. Recording by the exposure of a photo-sensitive surface to a signal-controlled light beam or spot.

1F24. Electrochemical Recording. Recording by means of a chemical reaction brought about by the passage of signal-controlled current through the sensitized portion of the record sheet.

1F25. Electrolytic Recording. That type of electrochemical recording in which the chemical change is made possible principally by ionization.

1F26. Electrothermal Recording. That type of electrochemical recording in which the chemical change is produced principally by thermal action.

1F27. Electromechanical Recording. Recording by means of a signal-actuated mechanical device.

1F28. Carbon-Pressure Recording. That type of electromechanical recording in which a pressure device acts upon carbon paper to register upon the record sheet.

1F29. Ink-Vapor Recording. That type of electromechanical recording in which vaporized ink particles are directly deposited upon the record sheet.

1F30. Facsimile Recorder. The part of the facsimile receiver in which the picture signal in its final form is systematically registered upon a record sheet as a facsimile of the subject copy.

1F31. Recording Spot. The instantaneous area impressed by the registering system of the recorder.

NOTE: The width is measured in the direction of line scanning; height, across this direction.

1F32. Synchronizing. The maintenance of predetermined speed relations between the scanner and the recorder within each scanning line.

1F33. Frame. A rectangular area, the width of which is the available line and the length of which is determined by the service.

1F34. Framing. The adjustment of the picture to a desired position in the direction of line progression.

1F35. Framer. A device for adjusting the equipment so that the recorded elemental area bears the same relation to the record sheet as the corresponding transmitted elemental area bears to the subject copy in the direction of line progression.

1F36. Phasing. The adjustment of the picture position along the scanning line.

1F37. Phaser. A device for adjusting the equipment so that the recorded elemental area bears the same relation to the record sheet as the corresponding transmitted elemental area bears to the subject copy in the direction of the scanning line.

1F38. Phasing Line. That portion of the length of scanning line set aside for the phasing signal.

NOTE: The phasing line may be expressed as a percentage of the length of scanning line.

1F39. Reproduction Speed. The area of copy recorded per unit time.

1F40. Nominal Line Width. The reciprocal of the number of lines per unit length in the direction of line progression.

1F41. Underlap. The amount by which the effective height of the scanning spot falls short of the nominal width of the scanning line.

NOTE: When using a rectangular spot, underlap may be expressed as a percentage of the nominal width of scanning line.

1F42. Overlap. The amount by which the effective height of the scanning spot exceeds the nominal width of the scanning line.

NOTE: When using a rectangular spot, overlap may be expressed as a percentage of the nominal width of scanning line.

1F43. Jitters. The distortion in the received picture caused by momentary errors in synchronism between scanner and recorder.

1F44. Skew. The deviation of the received frame from rectangularity due to asynchronism between scanner and recorder. Skew is expressed numerically as the tangent of the angle of this deviation.

SECTION 2. TRANSMISSION

2F1. Light Modulation. The method of introducing the carrier by periodic variation of the scanner light beam, the average amplitude of which is varied by the density changes of the subject copy.

2F2. Electrical Modulation. The method in which the carrier is introduced into an electrical modulator together with the signal currents directly produced by the density changes of the subject copy.

2F3. Positive Modulation. In an amplitude-modulation system, that form of modulation in which the maximum transmitted power corresponds to the minimum density of the subject copy. In a frequency-modulation system, it is that form of modulation in which the highest transmitter frequency corresponds to the minimum density of the subject copy.

2F4. Negative Modulation. In an amplitude-modulation system, that form of modulation in which the maximum transmitted power corresponds to the maximum density of the subject copy. In a frequency-modulation system, it is that form of modulation in which the highest transmitter frequency corresponds to the maximum density of the subject copy.

2F5. Subcarrier. An intermediate wave modulated by the facsimile signals and in turn used to modulate the main carrier, either alone or in conjunction with subcarriers on other channels.

NOTE: Its frequency is usually low relative to that of the main carrier.

2F6. Converter. A device which changes the type of modulation delivered by the scanner.

2F7. Maximum Keying Frequency (or Maximum Modulating Frequency). In a facsimile system, the

frequency in cycles per second numerically equal to one half the number of critical areas of the subject copy scanned per second.

2F8. Nominal Band. The frequency band of a facsimile-signal wave equal in width to that between zero frequency and the maximum keying frequency.

NOTE: The frequency band occupied in the transmitting medium will in general be greater than the nominal band.

2F9. Vestigial Sideband. The transmitted portion after a sideband has been largely suppressed by a transducer having a gradual cutoff in the neighborhood of the carrier frequency, the other sideband being transmitted without substantial suppression.

2F10. Vestigial-Sideband Transmission. That method of signal transmission in which one normal sideband and the corresponding vestigial sideband are utilized.

2F11. Facsimile-Signal Level. An expression of the maximum signal power or voltage created by the scanning of the subject copy as measured at any point in a facsimile system.

NOTE: According to whether the system employs positive or negative modulation, this will correspond to picture white or black, respectively. It may be expressed in decibels with respect to some standard value such as 1 milliwatt or 1 volt.

2F12. Peak-Signal Level. An expression of the maximum instantaneous signal power or voltage as measured at any point in a facsimile system.

NOTE: This includes auxiliary signals

2F13. Picture White (or "White"). The signal at any point in a facsimile system produced by the scanning of a selected area of subject copy having minimum density.

2F14. Picture Black (or "Black"). The signal at any point in a facsimile system produced by the scanning of a selected area of subject copy having maximum density.

2F15. White-to-Black Amplitude Range. In a positive amplitude-modulation facsimile system, the signal-voltage or -current ratio of picture white to picture black at any point in the system. In a negative amplitude-modulation system, it is the signal-voltage or -current ratio of picture black to picture white.

NOTE: This ratio is often expressed in decibels.

2F16. White-to-Black Frequency Swing. In a frequency-modulation facsimile system, the numerical difference between the signal frequencies corresponding to picture white and picture black at any point in the system.

2F17. Compression. A reduction in white-to-black amplitude range or frequency swing occurring between two points in the system.

2F18. Expansion. An increase in white-to-black amplitude range or frequency swing occurring between two points in the system.

2F19. Overthrow (or Overshoot) Distortion. The distortion resulting when the maximum amplitude of the signal wave front exceeds the steady-state amplitude of the signal wave.

2F20. Underthrow Distortion. The distortion resulting when the maximum amplitude of the signal wave front is less than the steady-state amplitude which would be attained by a prolonged signal wave.

2F21. Tailing (or Hangover). The excessive prolongation of the decay of the signal wave tail.

2F22. Facsimile Transient. The more-or-less well-defined damped oscillation superimposed on the signal envelope which may develop as a result of distortion when a signal undergoes a sharp amplitude change.

NOTE: Under specific conditions a facsimile transient may appear as overthrow or underthrow distortion or as tailing.

SECTION 3. GENERAL

3F1. Modulation. The process in which the amplitude, frequency, or phase of a wave is varied with time in accordance with a signal.

3F2. Modulated Wave. A wave of which either the amplitude, frequency, or phase is varied in accordance with a signal.

3F3. Carrier Wave. A wave generated at a point in the transmitting system and modulated by the signal.

3F4. Dual Modulation. The process of modulating a common carrier wave or subcarrier by two different types of modulation (e.g., amplitude and frequency modulation), each conveying separate information.

3F5. Channeling. The utilization of a modulation-frequency band for the simultaneous transmission from two or more communication channels in which the separation therebetween is accomplished by the use of carriers or subcarriers, each in a different discrete frequency band forming a subdivision of the main band.

NOTE: This covers a special case of multiplex transmission. See IT17 in "Standards on Transmitters and Antennas, 1938."

3F6. Multipath Transmission (or Multipath). The propagation phenomenon which results in signals reaching the radio receiving antenna by two or more paths, usually having both amplitude and phase differences therebetween.

NOTE: In European practice this is called "echo," with which term (3F9, below), compare.

3F7. Fading. The variation in intensity of radio

signals resulting from changes in the transmission medium.

3F8. Selective Fading. Fading in which the variation of radio field intensity is not the same at all frequencies in the frequency band of the received wave.

3F9. Echo. A wave which has been reflected at one or more points in the transmission medium, with sufficient magnitude and time difference to be perceived in some manner as a wave distinct from that of the main transmission.

3F10. Multipath Cancellation. The occurrence of effectively complete cancellation of the signals because of the relative amplitude and phase differences of the components arriving over the separate paths.

NOTE: This term is usually used in connection with the square-wave modulation of the radio-frequency carrier or of a subcarrier and generally results in the recording of only the starting and ending transients to the exclusion of the steady-state energized condition.

3F11. Elongation. The extension or elongation of the envelope of a signal due to the delayed arrival of certain of the multipath components.

3F12. Signal-Wave Envelope. The contour of a signal wave which is composed of a series of wave cycles.

3F13. Wave Front. Of a signal-wave envelope, that part (in time or distance) between the initial point of the envelope and the point at which the envelope reaches its crest.

3F14. Wave Tail. Of a signal-wave envelope, that part (in time or distance) between the termination of the steady-state value (or crest when the steady-state value is not reached) and the end of the envelope.

3F15. Power Level. An expression of the power being transmitted past any point in a system.

3F16. Transmission Level. The ratio of the signal power at any point in a transmission system to the power at some point in the system chosen as a reference point. This ratio is usually expressed in decibels.

3F17. Transmission Loss (or "Loss"). A general term used to denote a decrease in power in transmission from one point to another. Transmission loss is usually expressed in decibels.

3F18. Transmission Gain (or "Gain"). A general term used to denote an increase in power in transmission from one point to another. Transmission gain is simply a negative transmission loss if both are expressed on the same logarithmic scale. Transmission gain is usually expressed in decibels.

3F19. Transfer Ratio. From one point to another in a transducer at a specified frequency, the complex ratio of the generalized force or velocity at the second point to the generalized force or velocity applied at the first point.

NOTE: Generalized force or velocity includes not only mechanical quantities but also other analogous quantities such as acoustical and electrical. The electrical quantities are usually electromotive force and current.

3F20. Transfer Admittance. From one pair of terminals of an electrical transducer to another pair, at a specified frequency, the complex ratio of the current at the second pair of terminals to the electromotive force applied between the first pair, all pairs of terminals being terminated in any specified manner.

3F21. Attenuation-Frequency Distortion (or Attenuation Distortion or Amplitude-Frequency Distortion). That form of wave distortion in which the relative magnitudes of the different frequency components of the wave are changed.

3F22. Nonlinear Distortion. That form of distortion which occurs when the ratio of voltage to current, using root-mean-square values (or analogous quantities in other fields), is a function of the magnitude of either.

3F23. Delay Distortion. That form of wave distortion in which the relative delays (either phase or envelope) of the different frequency components of the wave are changed.

3F24. Phase Delay. In the transfer of a single-frequency wave from one point to another in a system, the time of delay of a part of the wave identifying its phase.

NOTE: A peak or an intercept of the wave may be taken as identifying its phase. The phase delay is equal to the ratio of the phase shift to the angular frequency.

3F25. Envelope Delay. In the transfer of an amplitude-modulated wave from one point to another in a system, the time of delay of the envelope of the wave.

NOTE: If the system distorts the envelope, the envelope delay at a specified frequency is still defined with reference to a modulated wave which occupies a frequency bandwidth approaching zero. The envelope delay is equal to the slope of phase shift plotted against angular frequency.

3F26. Corrective Network (Shaping Network). An electrical network designed to be inserted in a circuit to improve its transmission properties, its impedance properties, or both.

3F27. Shaping Network. See "Corrective Network, 3F26."

3F28. Impedance Compensator. A device designed to be associated with a transducer for the purpose of giving the impedance of the combination a desired characteristic with frequency over a desired frequency range.

3F29. Attenuation Equalizer. A device for altering the total transmission loss of a circuit for various frequencies in order to make substantially equal the total transmission loss for all frequencies within a certain range.

3F30. Delay Equalizer. A device included in a transducer which is designed to make the phase angle of the transfer ratio of the transducer substantially linear with the frequency within a desired range, thus making the envelope delay substantially constant in that frequency range.

3F31. Noise. Any extraneous electrical disturbance tending to interfere with the normal reception of a transmitted signal.

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